THE ELECTRONICS MAGAZINE WITH THE PRACTICAL APPROACH

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Elektor Electronics

MIDI KEYBOARD

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Elektor Electronics

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Number 168

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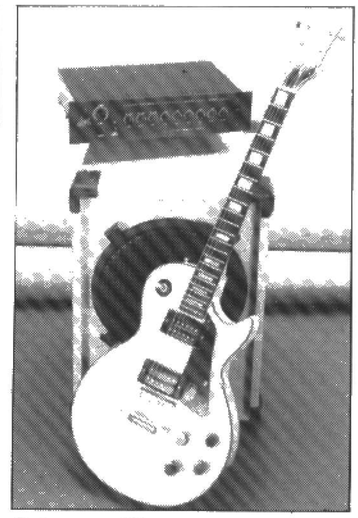
Front cover

Perhaps the major feature of the MIDI-compatible keyboard controller published in this issue is that it can be used with practically any existing keyboard, whether salvaged from a discarded instrument, or still in use in a piano, organ, or non-MIDI synthesizer. It supports up to 96 keys covering 8 octaves.



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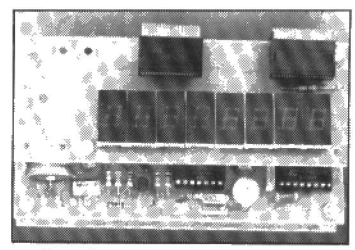

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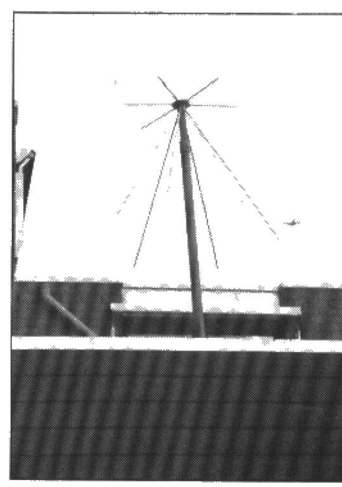

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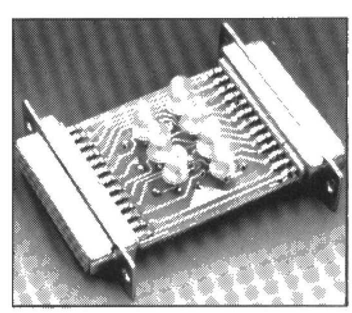

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EUROPEAN DIRECTIVE ON EMC

This magazine, in common with several others, has on several occasions in the past carried reports and news items on the rapidly and worryingly increasing interference caused by electromagnetic radiation from all kinds of electrical/electronic equipment, and that as a consequence the European Commission was planning legislation to curb such radiation. Appropriately, organizations such as ERA Technology[1] have been running seminars on electromagnetic compatibility for some time.

None the less, according to the Department of Trade and Industry, many electrical/electronic equipment manufacturers are not aware of either the European Commission's *Directive on ElectroMagnetic Compatibility* (EMC) or of its implications for them. The department is now preparing a consultative document to advise companies how to deal with the multiplicity of new technical standards and guidelines.

The warning bells from Whitehall to the industry toll at a time when trade figures for 1988 show that the electrical/electronics industry has reached an all-time high trade deficit of almost £4 billion. All sectors of the industry showed calamitous increases in trade deficit compared with 1987. To comply with the new directive, the industry is faced with additional costs of production, which further adversely affect our export performance.

Fortunately, although the directive was planned originally to become effective on 1 January 1990, it has been put back two years owing to the difficulties of getting the necessary standards ready in time. This postponement is bound to meet with sighs of relief in many a boardroom or chief engineer's office. None the less, it is of vital importance that the directive is taken into consideration in all designs without delay.

The directive, which is a vital component of the proposed single European market, is aimed at ensuring reliable and safe operation of all electrical and electronic goods throughout the community. It therefore contains a number of objectives for all electrical/electronic equipment of whatever design. Its intention is to advise manufacturers and designers on how to minimize radiation from a product as well as how to protect it from spurious emissions from outside sources.

It is understood that it is incumbent on manufacturers to comply with the objectives, not with a particular standard (do standards contain deficiencies?). Companies who feel that they may not be able to cope with the requirements of the directive may obtain help from the DTI or from organizations such as ERA Technology or the University of York's Electronics Centre[2], which have long-standing experience in dealing with problems associated with electromagnetic radiation.

1. ERA Technology Ltd • Cleeve Road LEATHERHEAD KT22 7SA • Telephone (0372) 374151
2. University of York •Heslington • YORK YO1 5DD • Telephone (0904) 430000

Ulster – 1992 Communications Bridgehead Between Europe and the United States

Northern Ireland is to have what is claimed to be the most advanced telecommunications system in Europe under a £100 million investment plan entered into by British Telecom, the UK Department of Economic Development and the European Commission.

The system is partly funded under the EC's *Special Telecommunications Action for Regional Development* (STAR) programme.

According to a BT spokesman, the system will make distance irrelevant to business and, from 1992, should shift Northern Ireland from the edge of Europe into becoming a vital communications bridgehead between Europe and the United States.

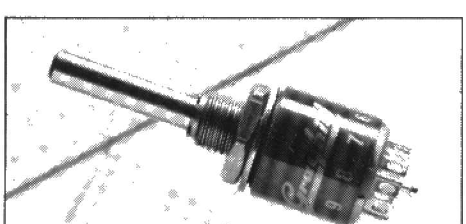
Direct Mobile to Mobile Communications in Prospect

Satellite manufacturers have been asked by the International Maritime Satellite Organization (INMARSAT) to suggest designs and technologies that will enable a new third generation satellite system to include such facilities as direct mobile to mobile communications between aeroplanes, ships and road vehicles.

1991 Start for UK Stereo Sound TV

The BBC is to start a UK stereo television sound broadcasting service in late 1991. The aim is that the service will be available to 60–70% of the population, provided they have suitably equipped TV receivers and video recorders.

Test transmissions of the Nicam 728 digital stereo system, developed by BBC engineers, which started in July last year, will continue with programme sound, originated in mono or stereo, on the digital stereo channel.



The Grayhill Series 50/51 rotary switch is available from Highland Distribution • Albert Drive • Burgess Hill • West Sussex RH15 9T

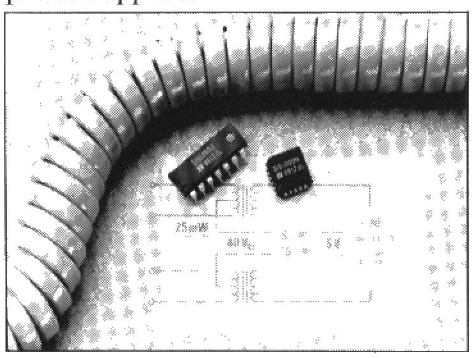
ELECTRONICS SCENE

New Marine VHF Channels to Ease Radio Congestion

The Department of Trade and Industry has assigned additional VHF radio frequencies for use by marinas, yacht clubs and pleasure craft to reduce congestion on the existing communications channel in coastal waters. The new channels will require a licence or, in the case of current (Channel M) licences, an amendment to this licence.

Switch-mode Power IC from Siliconix

Siliconix has introduced the world's first monolithic IC (Type Si9105) to meet the CCITT requirements for ISDN power supplies.



The new IC is a high-voltage switch-mode designed for low-power, high-efficiency DC/DC convertors. It replaces about a dozen discrete components.

USSR Radio Contract for Marconi

Marconi Communication Systems has won a contract from the USSR to supply a VHF coast station communication system that will provide traffic control of merchant shipping approaching and using the Gulf of Nakhodka.

Optical Fibre Test Instruments from STC Instrument Services

STC Instrument Services has introduced a range of optical fibre test instruments designed for testing communication networks in a variety of applications such as telecommunication systems, local area networks, computer installations, industrial process controls and high-voltage monitoring circuits.

The power meters, OTP610 and OTP620, are capable of measuring light output at 850, 1300 and 1550 nm, offering a test facility for the complete range of fibre optic networks in use.

Incorporation of a processor to compute the output on the clear LCD allows for automatic compensation of thermal errors as well as permitting the special range of measurements required when testing fibre optic cables up to 1 mm in diameter.

For details, contact Paul Channell at STC Instrument Services • Dewar House • Central Road • Harlow • Essex CM20 2TA.

Licensing of Low-power Radio Devices Abolished

As from last month, there is no need for licences for a wide range of low-power radio devices such as garage door openers, certain types of burglar alarms, industrial remote control equipment, radio microphones, low-power microwave devices and children's "walky-talkies".

The exempted categories are set out in an information sheet that is available from: DTI • Radiocommunications Division Library • Room 605 • Waterloo Bridge House • Waterloo Road • LONDON SE1 8UA.

Low-cost GaAs FET from Avantek

Avantek has introduced a new low-cost, low-noise GaAs FET suitable for C and Ku band earth station preamplifiers and other microwave applications. Designated AFT-13284, the transistor is useful as a low-noise amplifier over the 2–16 GHz range, and as an oscillator at up to 25 GHz. It has a noise figure of 0.7 dB with 15 dB associated gain at 4 GHz, 1.6 dB with 8 dB gain at 12 GHz.

Data sheets are available from Avantek distributors or direct from Avantek • M/S M82 • 481 Cottonwood Drive • Milpitas CA95035 • USA.

IEE Benelux

With a view to 1992 and the 'Single European Market', the Institute of Electrical Engineers has opened an overseas branch for Benelux, where 364 members of the institute are living. Once established, the group will be holding seminars, symposia, etc. on a wide range of topics related to

the electrical and electronics industry. For further information, please contact Michael Chadwick • Dept. XRI • ESTEC Postbus 299 • 2200 AG Noordwijk • The Netherlands.

25,000 Simultaneous Phone Calls on One Fibre

The equivalent of 25,000 simultaneous telephone conversations have been carried out over a single optical fibre link in British Telecom's network in a record-breaking demonstration of a technique that offers even bigger increases in capacity in the future.

The demonstration was carried out on a fibre in the optical submarine cable between the Cumbrian coast and the Isle of Man. The system, which came into operation last summer, operates without regenerators over its entire 94 km length.

British Telecom is the first to use optical wavelength division multiplexing over its operational network by sending light at different colours (frequencies) simultaneously along the same hair-thin optical fibre.

New Frequencies for Theatre Radio Microphones

An extra 22 radio frequency channels have been made available to theatres and concert halls that use radio microphones in a bid to overcome outside interference.

Manufacturers, retailers and suppliers of equipment, and theatre managements, will need to switch to the new approved frequencies and equipment as soon as possible. An information sheet giving details of the new and existing arrangements for authorized use of such radio microphones is available free of charge from: DTI • Radiocommunications Division Library • Room 605 • Waterloo Bridge House • Waterloo Road • LONDON SE1 8UA.

Communications Network of 21st Century

British Telecom's blueprint for the communications network of the 21st century—optical fibre "pipelines" carrying stereo television, high-fidelity stereo radio, telephone calls, information technology and other interactive services—becomes reality from next year.

About 500 business and residential customers at Bishops Stortford

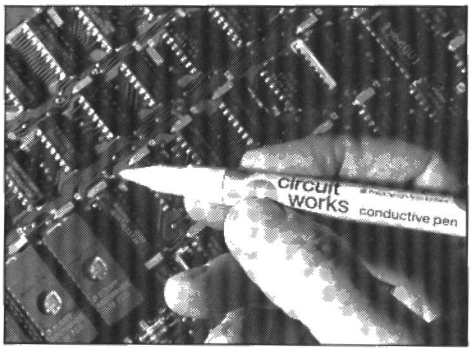
ELECTRONICS SCENE

are to take part in a trial that will enable BT to compare and contrast the operation and cost of two entirely different fibre systems its research engineers have developed. One is known as TOPN—telephony over passive networks. This is a low-cost system using a series of inert fibre splitters, or couplers, to 'siphon off' services to each customer on the network.

The other system is broadband integrated distributed star—BIDS— an active network using electronic switches to route to customer services they have selected. One of the most advanced systems of its kind in the world, its design will exploit experience gained in the British Telecom design of switched star networks used by the Westminster Cable TV franchise.

Electronic Tracing by Pen

Planned Products have introduced a low-cost pen that makes applying solderable electronic traces to most surfaces as easy as writing. The Circuit Works Pen, which writes in a highly conductive silver ink, has many applications including printed circuit board manufacture and repair, electromagnetic shielding and conductive point-to-point traces.

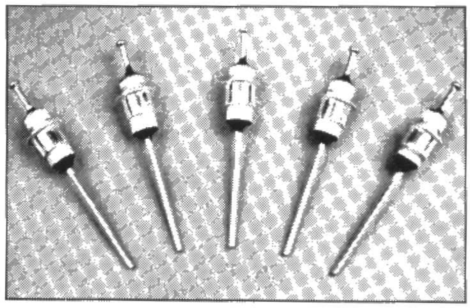


The pen incorporates a valved tip to allow the smooth application of the liquid silver conductor. Traces as narrow as 1/16 inch (1.6 mm) may be drawn. The pen comes filled with enough silver to make 150 ft (just over 45 m) of conductive traces.

Single pens sell at $9.95 plus $1.00 postage ($2.00 outside the USA) and handling and are available from Planned Products • 21105 Santa Cruz Highway • Brush Road • Dept FEE • Los Gatos • CA95030 • USA • Phone (408) 353–4251 • Fax (408) 354–4818.

EISA Token Ring

Proteon is developing an Extended Industry Standard Architecture (EISA) bus master Token Ring card that will operate at 4 Mbits/s and 16 Mbits/s. The product is expected to be ready this autumn to coincide with the first release of EISA-based computers.



Samples of the new Murata Erie Type 1214 feed-through EMI / RFI filters. These devices are available only from STC Mercator

Fast Growing Data Traffic

According to *Digital Telecommunications Market in Europe*, a new report from Frost & Sullivan, data traffic in western Europe is growing about six times as fast as voice traffic among business users. Already, Europeans invest close to £13 billion a year in digital communications equipment.

New Radio Paging Transceiver

The new 'Keyfon' radio paging transceiver from ANT Telecommunications will (1) receive verbal messages when calls are initiated through a company's PABX system via appropriate interface equipment or from a central control unit; (2) enable calls to be made to any other paging transceiver in the system; (3) allow users to make a call back to the central control unit or to any telephone handset; (4) allow calls to be made direct to the public telephone network via the user's own PABX system.

The 'Keyfon' is based on a modular 'building block' approach and provides facilities from tone bleep and display to full two-way speech. It is individually programmed with relevant information regarding each transceiver's identification number being held on an interchangeable code plug.

All enquiries to ANT Telecommunications Ltd • 17 Liverpool Road • Slough • Berkshire SL1 4QZ • Telephone (0753) 820242.

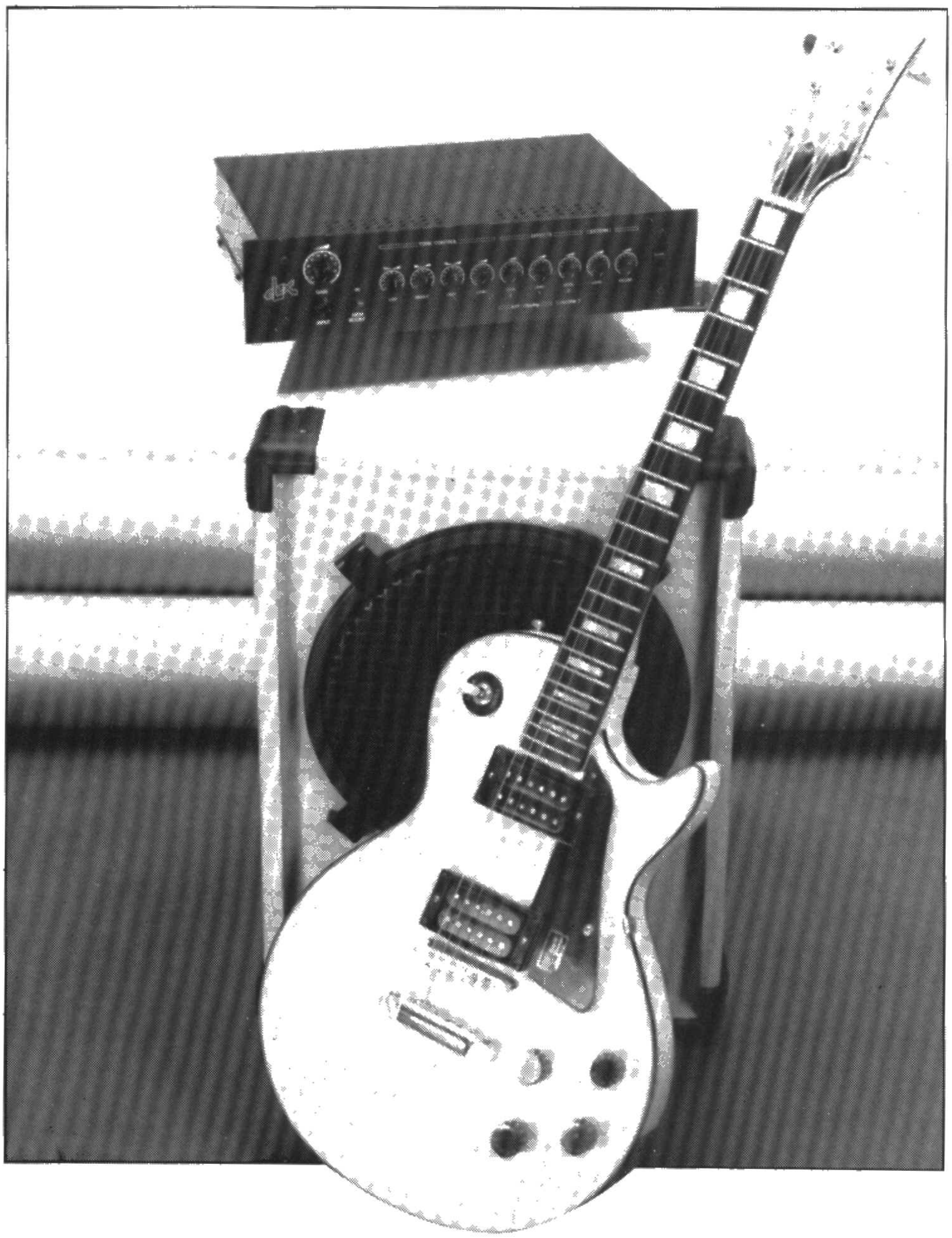
INTERMEDIATE PROJECT

A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary understanding and knowledge of electronic engineering. Each project in the series will be based on inexpensive and commonly available components.

4. Guitar amplifier

by T. Giffard

This modular guitar amplifier has features found on commercially available models well outside the financial reach of the novice or intermediate electronics enthusiast with a passion for playing the electric guitar. These features include tone control, two different fuzz units, a mixer, a level indicator and, last but not least, a cool sixty watts of output power obtained from a ready-made amplifier module.



The guitar amplifier forming this month's Intermediate Project is a design that will give many hours of enjoyment in building and, more importantly, using it. The overall design may appear fairly complex at first, but the individual modules are all fairly simple to build on the universal prototyping board used for nearly all projects in this series of articles. What's more, the modules are all based on low-cost components, and their operation is discussed in detail. To get an overall idea of what is required to build the amplifier, a quick overview is given of the individual modules.

Figure 1 shows the input amplifier built from discrete parts (i.e., there are no integrated circuits). The amplifier is followed by a tone control section and the first of two fuzz (distortion) circuits. Next follows the second fuzz circuit and associated level indicator (Fig. 3). Figure 4 shows the mixer, and Fig. 5, finally, the voltage regulator that forms part of the power supply. The mains transformer is not shown in Fig. 9 because it is contained in the power supply (PSU422) that must be purchased together with the HY128 amplifer module from ILP. Wiring diagram Fig. 6 shows the interconnections between the modules, and the power amplifier module Type HY128, a ready-made 60 W unit with superb specifications.

Preamplifier

A conventional preamplifier for audio signals is unsuitable for use with an electric guitar because it can not cope with the large amplitude range of the signal from the pick-up element. Also, the capacitance and impedance of the pick-up element

and the cable between the guitar and the amplifier are fairly troublesome parameters.

The preamplifier proposed here is designed specially for use with electric guitars, and has all the necessary characteristics to give good performance with almost any pick-up device.

The circuit diagram of Fig. 1 shows that the guitar signal is applied to FET T_1 via coupling capacitor C_1. The FET functions as an impedance converter. Potential divider R_2-R_3 sets the source voltage at about half the supply level. The guitar signal at the source is capacitively fed to a discrete differential amplifier formed by T_2, T_3 and T_4. These are Types BC550C to keep the overall noise level as low as possible. The base and collector resistors of T_2 and T_3 have relatively low values for the same reason, and necessitate the use of the source follower, T_1, without which the guitar element would be too heavily loaded.

The specifications of the amplifier are impressive given the simplicity of the design: at an amplification of 100 (adjustable with P_1), the differential stage has a signal-to-noise ratio (S/N) greater than 100 dB. This increases to 130 dB for amplifications between 1 and 30. Mind you: the stated amplification applies to the circuit *without* T_1: when the FET is included, the amplification is reduced to about 65.

As customary with preamplifiers, an output stage is provided for driving the tone control circuit. The drive margin of the output amplifier is determined mainly by the current sunk by current source T_5, and the load impedance at the output. Capacitor C_3 is connected across the base-collector junction of T_6 to prevent the parasitic capacitances around the transistor causing signal-dependent oscillation. The bandwidth of the preamplifier is limited to about 20 kHz with the aid of C_{12}. Bandwidth reduction is necessary to minimize the risk of distortion caused by RF interference at high amplification settings.

A feature of the preamplifier is the built-in 'high-boost' circuit composed of L_1, R_{11} and C_4. This L-C-R circuit is tuned to about 10 kHz, to present a relatively high impedance to low frequencies. This impedance becomes lower as the frequency increases: at 10 kHz, it is only slightly higher than 220 Ω, the value of R_{11}. The tuned circuit is connected in parallel with R_9, which forms part of feedback network P_1-C_{12}-R_9. The result of this configuration is that the voltage gain of the circuit at low frequencies is determined by the ratio P_1/R_9. The impedance of the L-C-R circuit drops with increasing frequency, and causes the amplification to rise accordingly. The additional gain at 10 kHz amounts to about 12 dB. A disadvantage of the high-boost circuit is that the L-C-R circuit becomes virtually ineffective when P_1 is set to minimum resistance. In that condition, the entire signal is fed to the output amplifier, which results in a fairly constant amplitude. In practice,

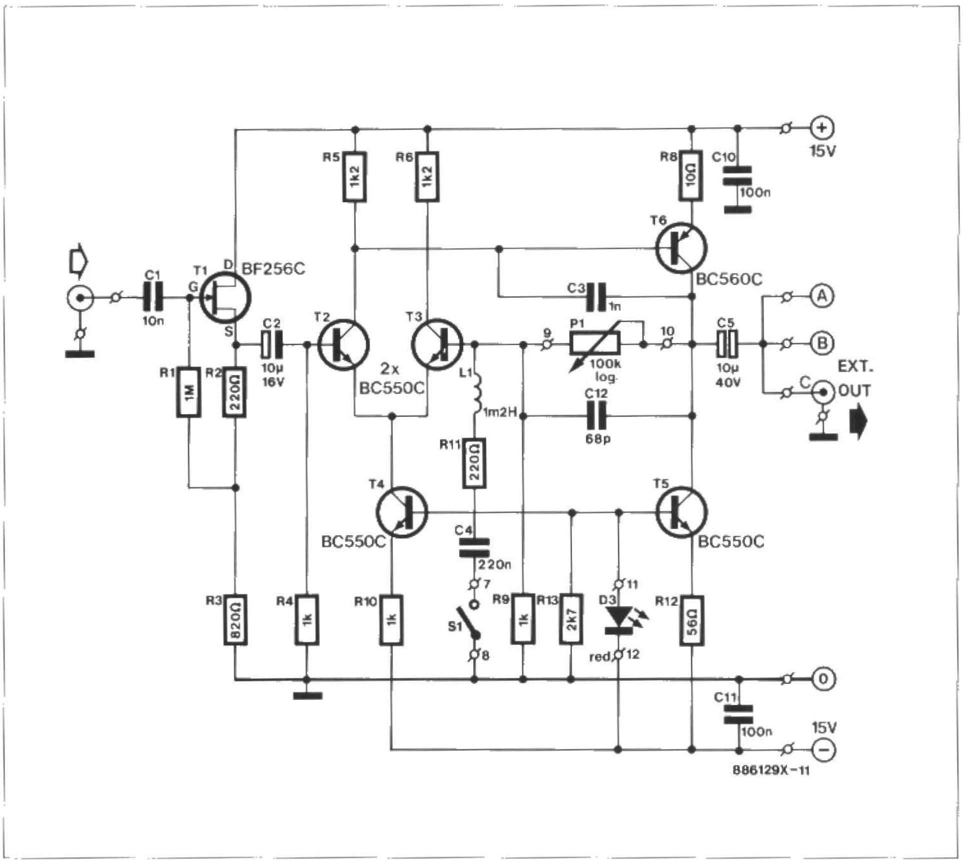


Fig. 1. The guitar preamplifier is a low-noise type based on a source follower and a differential amplifier.

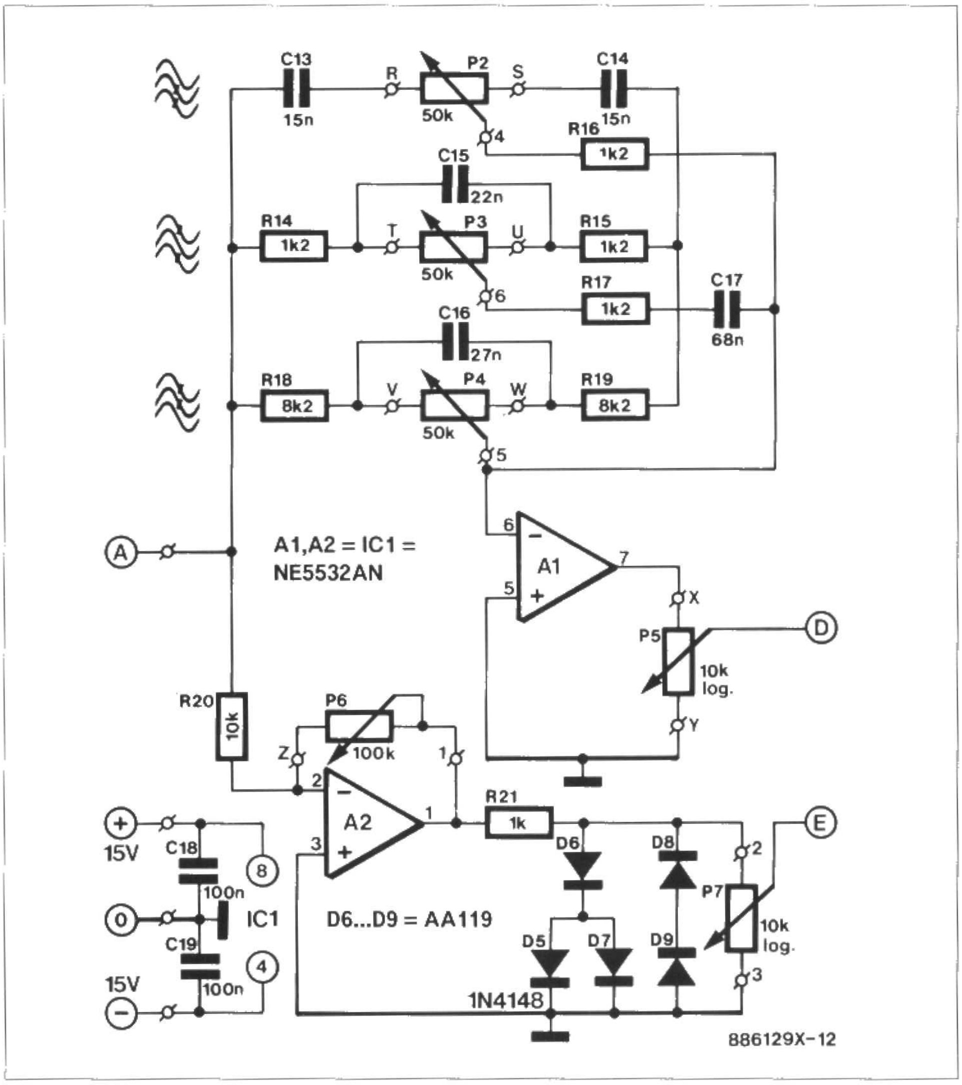


Fig. 2. Circuit diagram of the tone control and the first fuzz circuit.

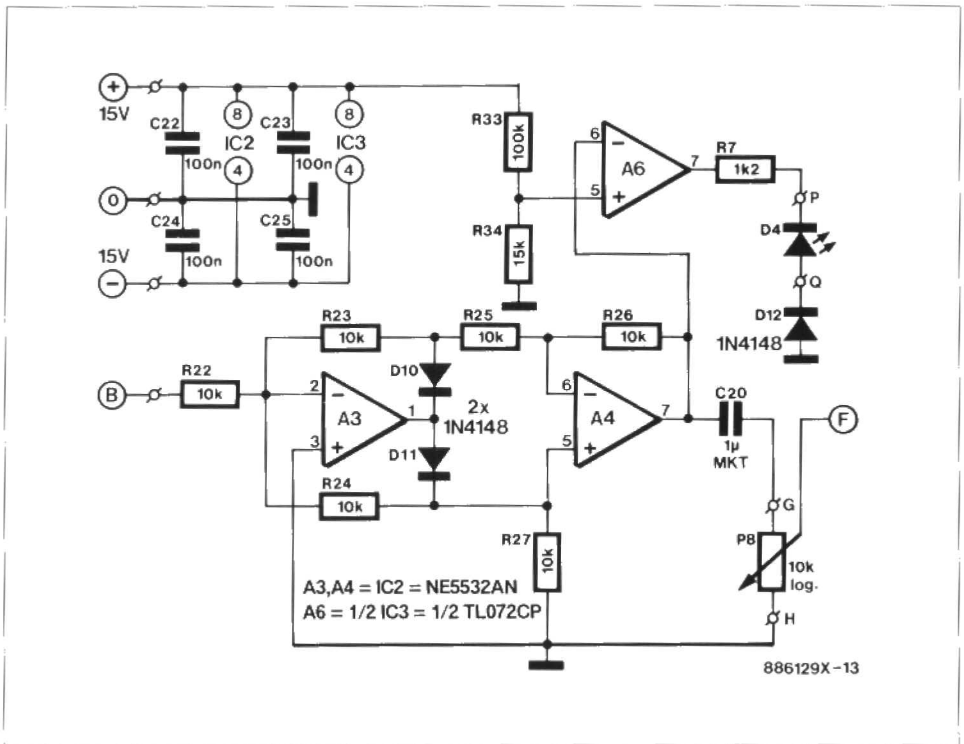


Fig. 3. Level indicator and second fuzz circuit.

this means that the high-boost effect is only obtained at relatively high gain settings.

Light-emitting diode (LED) D_3 does not, in principle, function as a visual indicator, but, together with R_{13}, as a reference for current sources T_4 and T_5.

Tone control and first fuzz circuit

The circuit diagram of Fig. 2 shows the tone control section around opamp A_1 which has 3 filter networks in the feedback path. Also shown is opamp A_2 and surrounding parts, which together form the first of two fuzz circuits, the so-called 'soft clipper'.

The tone control circuit is composed of 3 filter sections: a high-pass filter with a roll-off frequency of about 2 kHz (C_{13}, P_2 and C_{14}), a 'presence' (middle) control which is effective around 1 kHz (R_{14}, P_3, R_{15} and C_{15}), and a low-pass filter dimensioned for a roll-off frequency of about 200 Hz (R_{18}, P_4, C_{16} and R_{10}). The maximum amplification of the tone control section amounts to about 15 dB. Resistor R_{17} isolates the high and middle controls. As in the preamplifier, resistor values are kept as low as possible to keep noise down to an acceptable level. Fortunately, the preamplifier has sufficient drive to overcome the resulting, relatively low, input impedance of the tone control section.

The filtered signal is applied to P_5 via opamp A_1. The potentiometer allows setting the amount of filtered signal that is fed to the summing amplifier via point B.

The output signal provided by the preamplifier is also fed to the first fuzz circuit, set up around A_2. Controlled signal distortion is achieved with two anti-parallel connected pairs of germanium diodes, which provide the so-called soft-clipping effect. The sound obtained in this manner bears some resemblance to that of a typical valved guitar amplifier, but only if the input signal level is sufficiently high. Opamp A_2 sets the drive for the first fuzz circuit in conjunction with level control P_6. The higher the signal, the more effective the fuzzer. Care should be taken, however, not to overload A_2, since this would result in real clipping of the signal, which is undesirable. Fortunately, the use of germanium diodes results in a relatively high maximum input level of about 1 V_{rms}. Silicon diode D_5 raises the level of the second harmonic.

The distorted signal is fed to point E via level control P_7 (distortion level 1). From there, it is taken to the input with the same letter indication on the summing amplifier.

Second fuzz circuit and level indicator

The input of the second fuzz circuit is marked B in Fig. 3 to indicate that the module is connected to the second output of the preamplifier.

Two opamps, A_3 and A_4, two diodes, D_{10} and D_{11}, a coupling capacitor, C_{20}, and a handful of resistors together form a full-wave rectifier for audio signals. This circuit effectively inverts and inserts the negative half-waves of the input signal between the positive half-waves, doubling the input frequency. The output signal of A_4 is fed to input F of the summing amplifier via level control P_8 (distortion level 2), and to the – input of comparator A_6, which forms a level indicator in conjunction with D_4. The output signal of the full-wave rectifier is suitable for driving the level indicator because its amplitude is directly proportional to that of the output signal of the preamplifier. The output of A_6 carries a rectangular signal whose duty factor (pulse/pause ratio) is a measure of the guitar signal. The duty factor becomes smaller, and D_4 lights brighter, as the amplitude increases.

Summing amplifier

The circuit diagram of this module is given in Fig. 4. Opamp A_5 simply sums the signals applied to the 4 series resistors connected to the – input. The volume ratio of these signals is set by the relevant controls on the driving modules. An external input is also provided to allow the summing amplifier to be used as a mixer. Master volume control P_9 and coupling capacitor C_{21} feed the composite signal to the input of the ILP power amplifier module.

Power supply

As already mentioned, the supply voltage for the guitar is obtained from the power supply (PSU422) to be purchased with the ILP module. The symmetrical regulator shown in Fig. 5 provides the required ±15 V rails for feeding the previously discussed active circuits. Zener diodes D_1

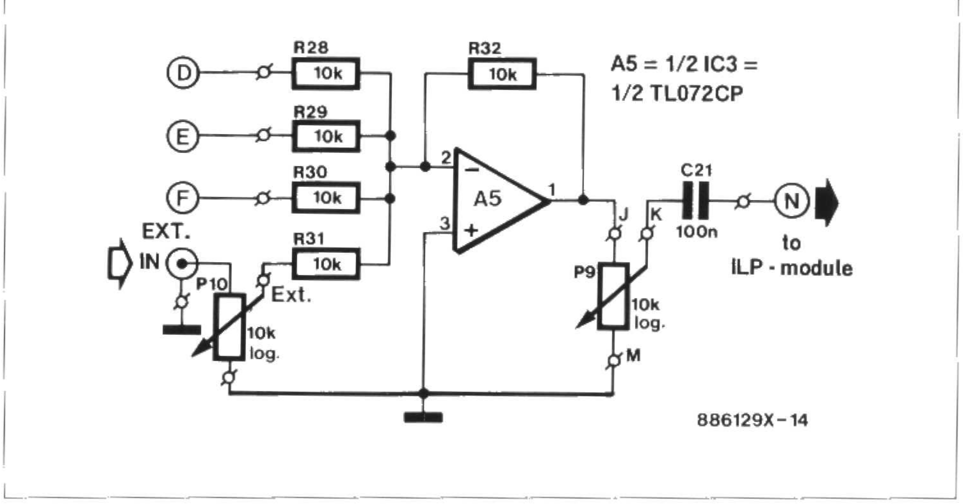


Fig. 4. Summing amplifier.

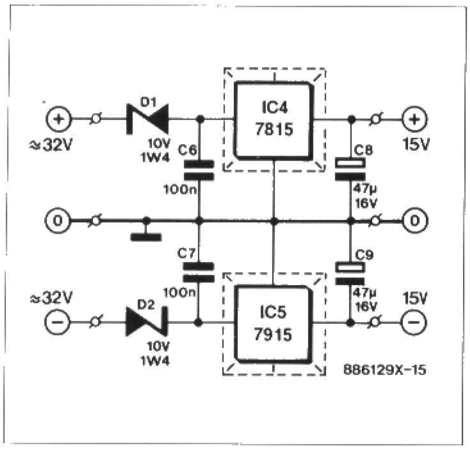


Fig. 5. Symmetrical regulator for stepping down the voltages taken from the PSU422.

and D_2 are connected ahead of the regulator inputs to limit the maximum input voltage to a safe level, and to reduce the dissipation in the 7815 and the 7915. None the less, both regulators should be fitted with TO220-style heat-sinks.

Construction time

After all this theory it is time to draw up the list of materials required for building the guitar amplifier. You will need:

- three universal prototyping boards UPBS-1 and the parts to fit on these;
- one ILP Type HY128 power amplifier module and Type PSU422 power supply,
- one 15-inch rack-type metal enclosure;
- one front panel to the design shown in Fig. 10;
- 10 potentiometers;
- input and output sockets;
- light-duty switches and one DPDT mains switch;
- tools;
- this article.

Start the construction by populating the prototyping boards as shown on the overlays in Figs. 6, 7 and 8. Each board has a fair number of soldering pins, which must be installed exactly as shown. Also do not overlook the wire links.

The completed circuit boards are put aside. The next step is the fitting of various parts into the enclosure. These parts include the transformer, the mains entrance socket with built-in fuse holder, the

Parts list

Resistors (±5%):
R_1=1M0
R_2;R_{11}=220R
R_3=820R
R_4;R_9;R_{10};R_{21}=1K0
R_5;R_6;R_{14}–R_{17} incl.=1K2
R_8=10R
R_{12}=56R
R_{13}=2K7
R_{18};R_{19}=8K2
R_{20};R_{22}–R_{32} incl.=10K
R_{33}=100K
R_{34}=15K
P_1=100K logarithmic potentiometer
P_2;P_3;P_4=50K linear potentiometer
P_5;P_7–P_{10} incl.=10K logarithmic potentiometer
P_6=100K linear potentiometer

Capacitors:
C_1=10n
C_2=10µ; 16 V
C_3=1n0 ceramic
C_4=220n
C_5=10µ; 40 V; bipolar
C_6;C_7;C_{10};C_{11};C_{18};C_{19};C_{21}–C_{25} incl.=100n
C_8;C_9=47µ; 16 V; tantalum
C_{12}=68p
C_{13};C_{14}=15n
C_{15}=22n
C_{16}=27n
C_{17}=68n
C_{20}=1µ0

Semiconductors:
T_1=BF256C
T_2–T_5 incl.=BC550C
T_6=BC560C
D_1;D_2= zener diode 10 V; 1.4 W
D_3;D_4= red LED
D_5;D_{10};D_{11};D_{12}=1N4148
D_6–D_9 incl.=AA119
IC_1;IC_2=NE5532AN
IC_3=TL072CP
IC_4=7815
IC_5=7915

Miscellaneous:
L_1=1mH2
S_1= miniature SPST switch.
Double-pole mains switch with integral lamp.
2 wander sockets.
2 phono or jack sockets.
1 jack plug.
1 mains entrance socket with integral fuse-holder.
1 fuse 500 mA; slow.
3 printed-circuit boards Type UPBS-1 (see Readers Services page).

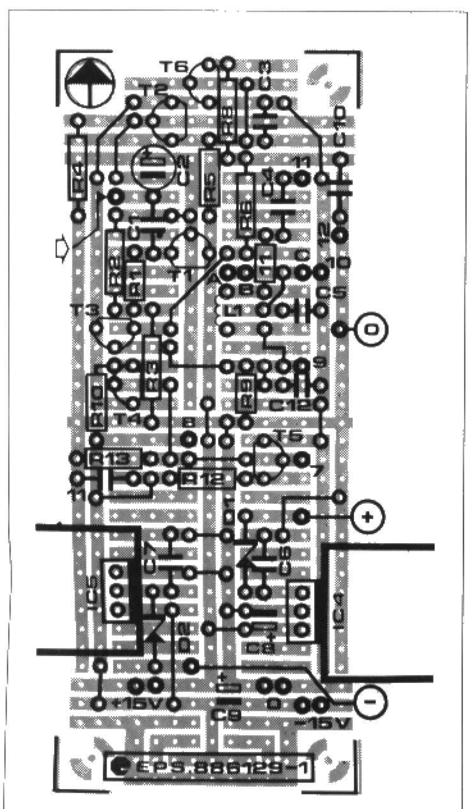


Fig. 6. Component mounting plan for the preamplifier board, which also houses the symmetrical 15 V regulator.

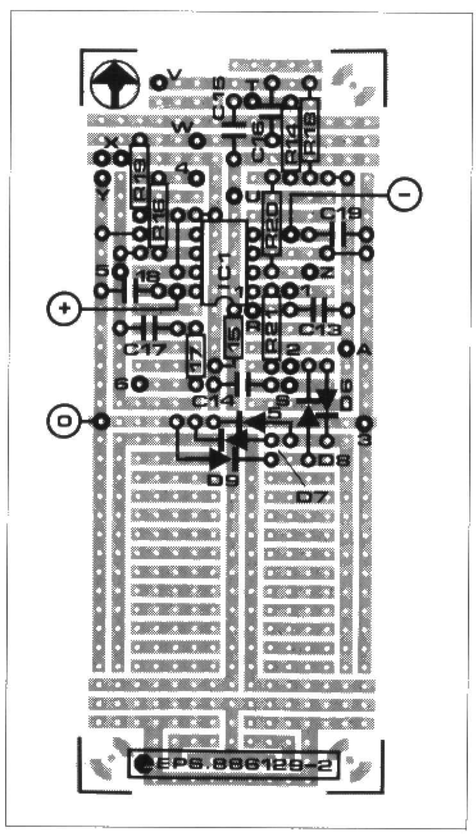


Fig. 7. Component mounting plan for the board that contains the tone contol and the first fuzz circuit.

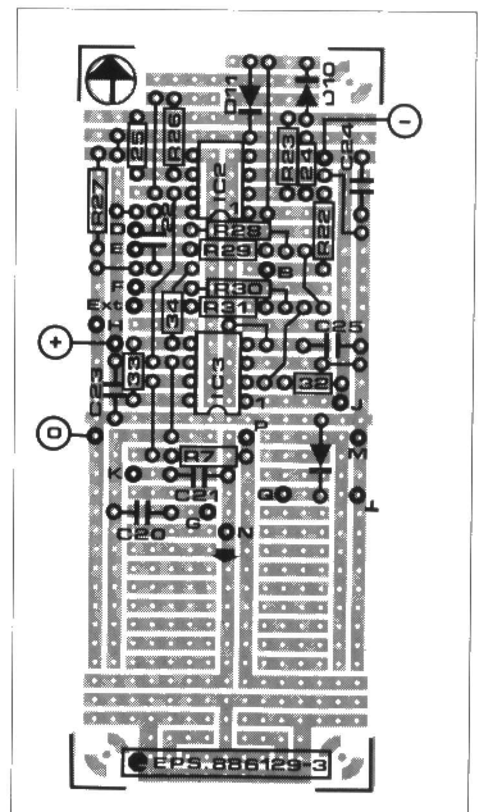


Fig. 8. Component mounting plan for the summing amplifier, the second fuzz circuit, and the level indicator.

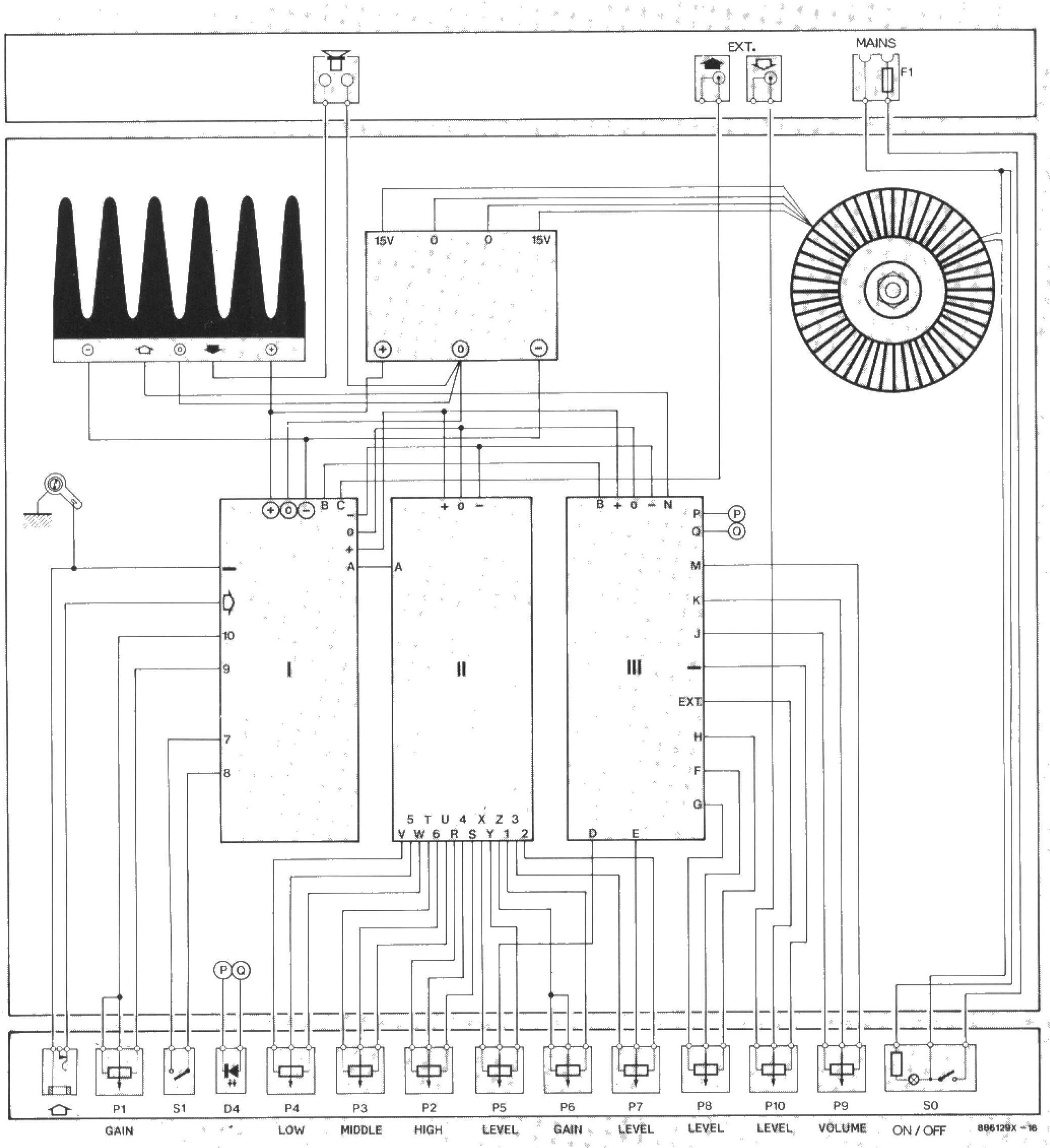


Fig. 9. Wiring diagram and suggested positions of the modules in the 15-inch rack enclosure.

power amplifier module, the three boards, the potentiometers, the input and output sockets, the level indicator LED, and the switches. Although the arrangement of the modules, the power amplifier and the power supply parts is not critical, it is recommended to follow the diagram given in Fig.9.

Make sure that the boards and the power supply section are well insulated from the enclosure. Plastic PCB spacers are ideal to achieve this.

Take your time to study and install the wiring as shown in Fig. 9. The power supply wires and the wires to and from the output sockets must have a cross-sectional area of at least 1.5 mm². The remaining connections may be made in light-duty wire, bearing in mind that the

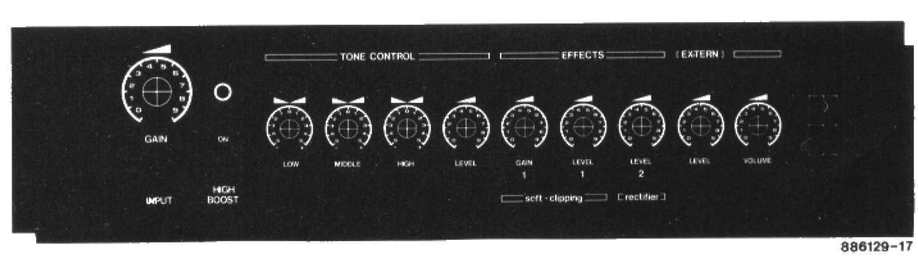


Fig. 10. Suggested front-panel lay-out.

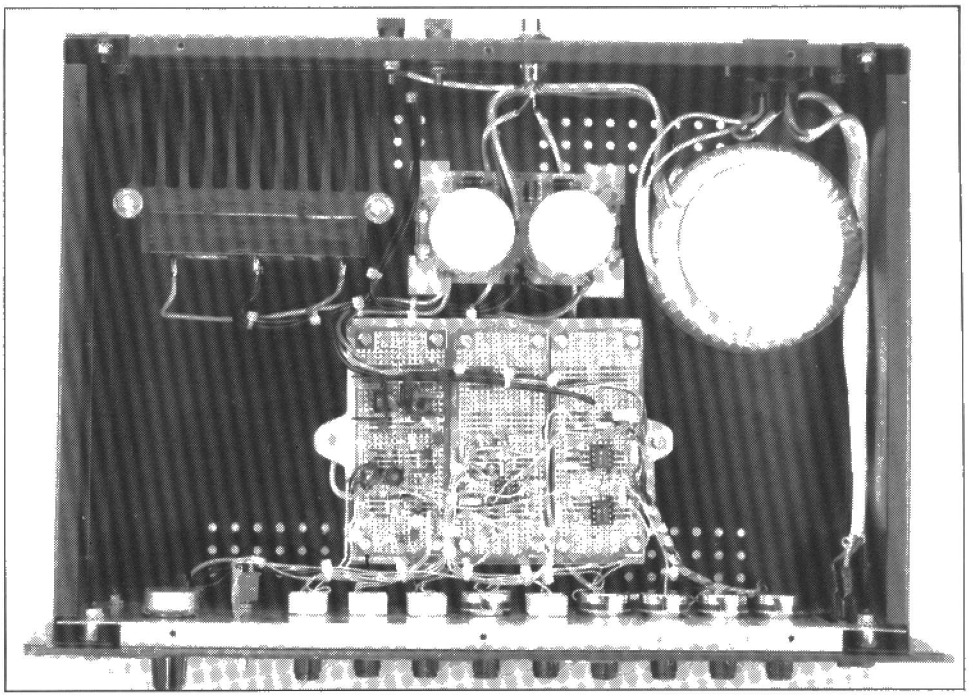
Table 1

Power amplifier HY128

• Output power:	60 W into 8 Ω
• frequency range:	15 to 15,000 Hz
• Harmonic distortion (1 kHz):	0.01%
• Signal-to-noise ratio:	100 dB
• Slew rate:	15 V/μs
• Loudspeaker impedance:	>8 Ω
• Input voltage:	500 mV
• Input impedance:	100 kΩ

Power supply PSU422

• Transformer:	toroid; 2×22 V
• Output voltage:	±32 V
• Output current:	1.5 A



wires to the potentiometers must be kept shorter than 5 cm. Longer wires inevitably pick up hum and noise. Where it is not possible to avoid relatively long distances, screened cables must be used for the signal connections. When screened cable is used, great care should be taken to keep the input and output sockets *insulated from the chassis*.

The power supply is a possible source of danger because of the mains voltage. Make sure that all the relevant parts are secured and well-insulated. Seek the assistance of an experienced engineer if you have not previously constructed mains-powered equipment.

Some noteworthy points

The use of a ready-made power amplifier module, the HY128 from ILP, avoids the problems involved in building a rugged, reliable 60 W amplifier with a large enough heat-sink. It is strongly recommended to purchase the HY128 complete with the associated toroidal mains transformer and power supply board, which contains a discrete bridge rectifier and two large electrolytic capacitors to provide the unregulated symmetrical supply voltage for the module. The main specifications of the set, which is available from Jaytee Electronic Services, are given in Table 1.

As already noted, the LED used in the preamplifier, D_3, functions as a reference device. Since it lights dimly, however, its function doubles into that of a power on/off indicator. The light intensity may not be high enough in all cases, so that a normal bulb may be preferred (the DPDT mains switch may have one inside).

Level indicator D_4 is, of course, to be fitted on to the front panel. If it lights continuously while playing, the amplification must be reduced by turning P_1 counter-clockwise.

Figure 9 shows that additional signal connections are provided on the guitar amplifier. One extra input socket is connected to potentiometer P_{10}, which sets the level of external effects units such as a wah-wah, a flanger, phaser or echo unit. The other socket is an auxiliary output which may be used to feed the preamplifier signal to other power amplifiers. The output from the tone control section may also be made available for other equipment: a soldering pin marked 'x' is provided for this purpose on the board.

CORRECTIONS

The digital model train (part 1)

March 1989, p. 52.

In some cases the operation of the locomotive decoder is affected by points control commands. This problem can be solved by increasing the value of R_1 from 12 kΩ to 39 kΩ. The circuit diagram (Fig. 16) should be amended accordingly.

LFA-150: a fast power amplifier (final part)

December 1988, p. 43.

On the component overlay of the protection board shown in Fig. 10, the plus sign at the negative pole of electrolytic capacitor C_{48} should be removed: the printed capacitor symbol indicates the correct polarity.

Autonomous I/O controller (part 1)

December 1988, p. 31.

Table 1 should be inverted: no diodes fitted gives instrument address 150-151, and both diodes fitted address 144-145.

Facsimile interface for Atari ST

January 1989, p. 15.

The pin connections for the two-wire RS-232 link given under the heading *Construction and alignment* should read:

pin 7: ground; pin 3: serial signal; connect pin 8 to pin 20 and pin 6; connect pin 4 to pin 5.

Pitch control for CD players

December 1988, p. 24.

On the component overlay of printed-circuit board 880165 (Fig. 7), the capacitor next to C_{21} should be marked C_{20}, not C_{19}. The value remains the same at 100 nF, but a ceramic capacitor should be used as advised in the Parts List.

Colour test-pattern generator

December 1988, p. 54.

Diodes D_{16}, D_{17} and D_{18} are shown with the wrong polarity on the component overlay (Fig. 5).

CMOS SWITCHES FOR AUDIO APPLICATIONS

T. Giffard

When about ten years ago the first analogue CMOS switches and multiplexers reached the audio components market, many audio enthusiasts believed that there was at last an end in sight to the use of expensive relays and other electromechanical elements to control volume and rumble or switch signal sources and functions. Unfortunately, the low speeds, high, non-linear on-resistance and level of crosstalk associated with the new devices soon put an end to these expectations. Over the past few years their quality is claimed to have improved considerably. These claims have been tested in our laboratory through a number of CMOS switches and circuits.

We will commence by taking from the numerous parameters of CMOS switches those that are of importance to audio designers, namely:

- resistance of the closed switch (R_{on} in Ω);
- analogue voltage range (U_a in V);
- R_{on} as a function of U_a (in %);
- consistency of R_{on} over a number of switches (in %);
- insulation in off condition (in dB);
- crosstalk between a closed and an open switch (C_T in dB);
- rise time (T_{on} in ns);
- drop-out time (T_{off} in ns).

The first four of these parameters are particularly important for the linearity of the audio circuit; the next two, for the crosstalk performance, and the rise time is of vital importance in some applications as we shall see later.

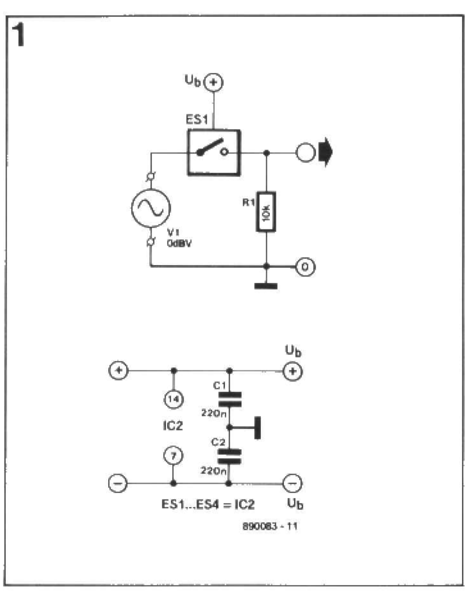


Fig. 1. This simple circuit is perfectly satisfactory for many audio applications.

Topology of CMOS switches

CMOS switches may be used for three specific functions: (1) the selection of the signal source; (2) switching of auxiliary functions, such as changing filter characteristics or altering the volume, in the same way as a rotary switch; and (3) as quasi-digital volume control.

In (1) and (2) the basic circuit of the switch is almost always the same: it serves to interrupt the signal path in a fairly simple manner. For instance, in

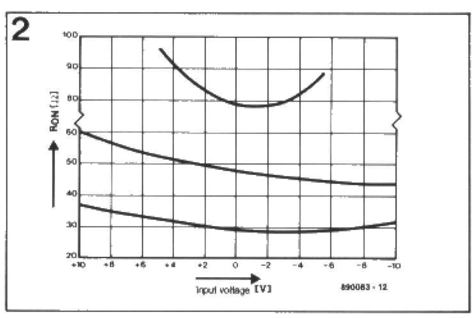


Fig. 2. On resistance vs supply voltage curves.

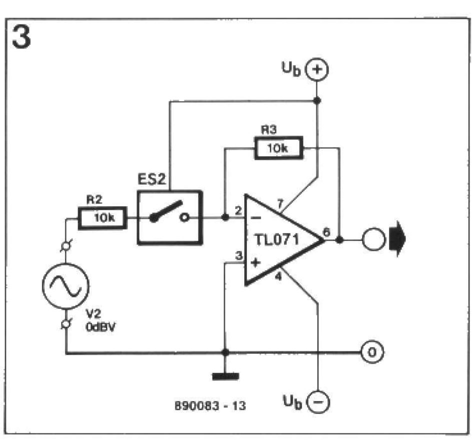


Fig. 3. An improved version of Fig. 1 for more exacting requirements.

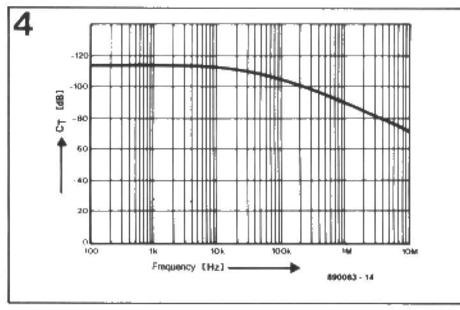


Fig. 4. Typical channel separation vs frequency characteristic.

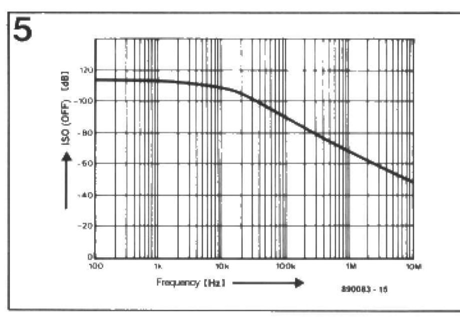


Fig. 5. Typical crosstalk vs frequency characteristic.

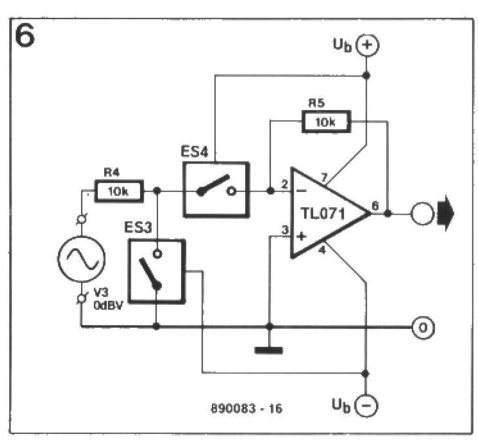


Fig. 6. An improved version of Fig. 3 for the most demanding applications.

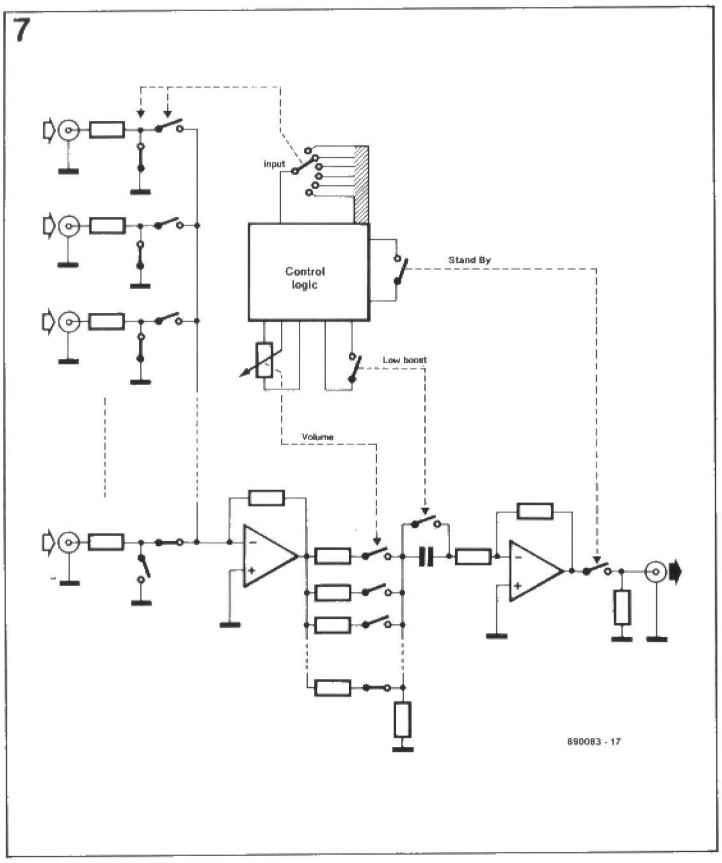


Fig. 7. Schematic diagram of a DC-controlled preamplifier.

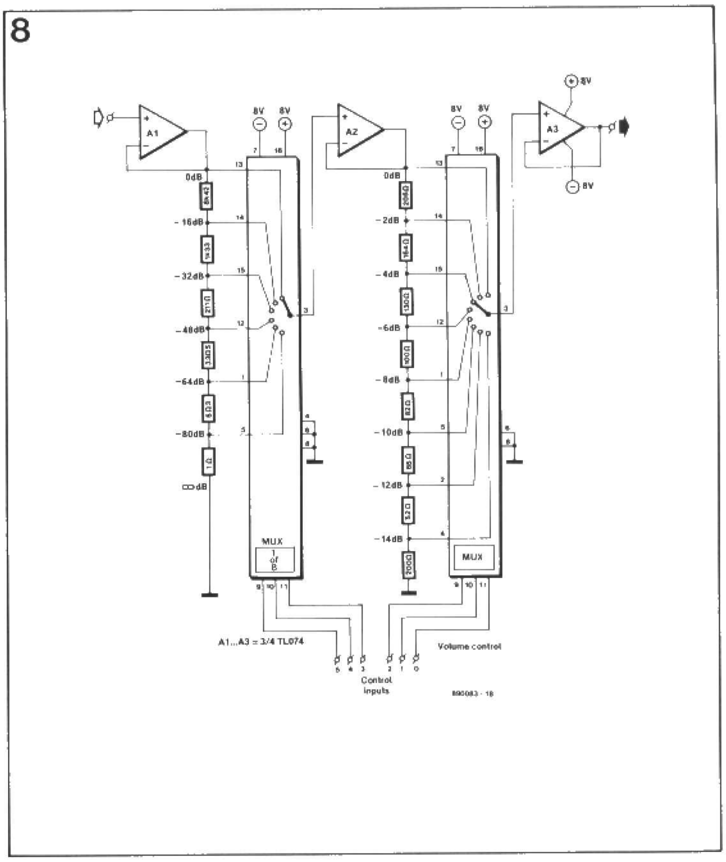


Fig. 8. Traditional high-quality electronic volume control covering a range of 96 dB in 2 dB steps.

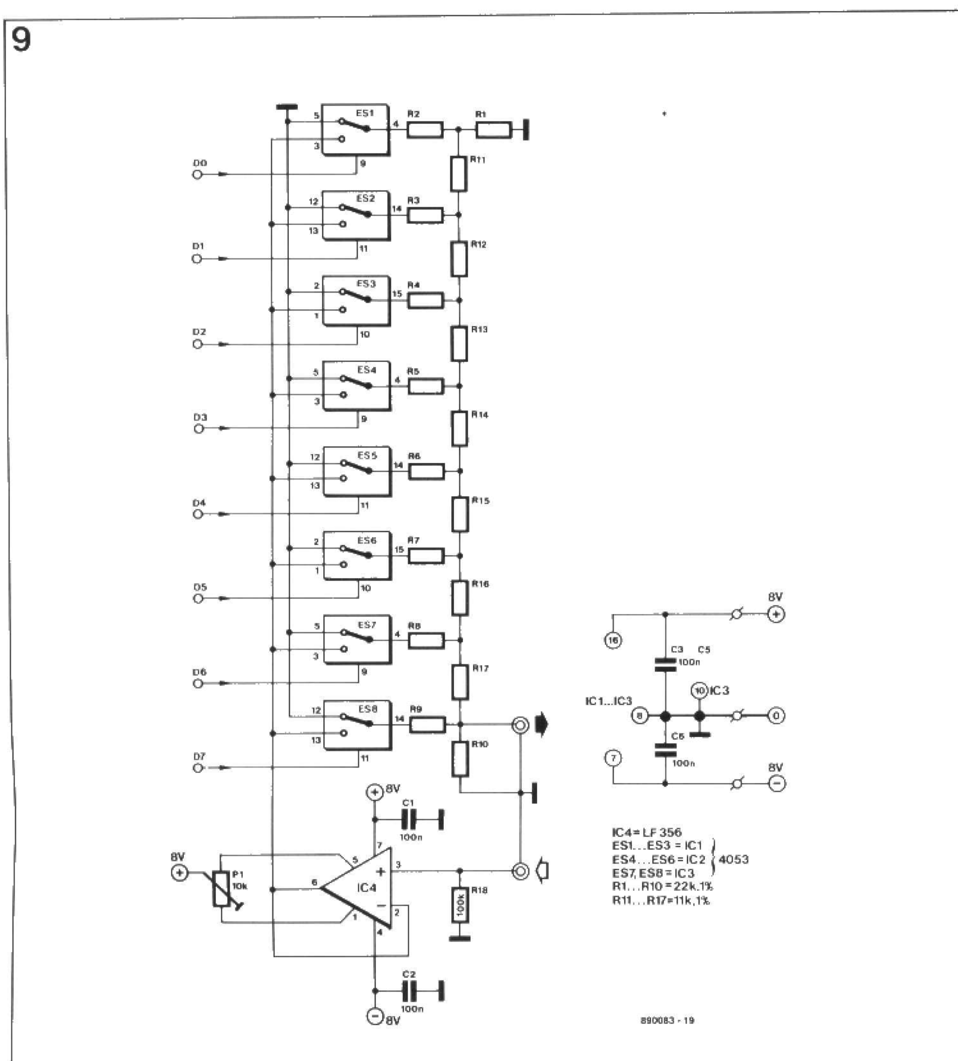


Fig. 9. Alternative to Fig. 8 with electronic step control via single CMOS switches.

Fig. 1 the popular CD4066 has been inserted into the signal path to serve as a relay or electromechanical switch. The 10 kΩ load resistance is part of a general audio network. Tests of this circuit were reasonably satisfactory in spite of the dependence of R_{on} on the signal level and supply voltage (typical curves of the former are given in Fig. 2). The relatively large value of R_{on} and that of the ratio R_{on}:R_l caused some distortion of the signal.

The tests also showed that CMOS switches, even from the same manufacturer, vary quite a lot from one to another.

The overall distortion varied from −74 dB to −84 dB (<0.02%), depending on the IC, at a supply voltage of ±7.5 V and a signal level of 1 V r.m.s. The distortion remained within the values indicated when the signal level was increased, but increased sharply when the supply voltage was reduced.

This IC can not be recommended for use in exacting applications, but for normal purposes it is perfectly satisfactory.

The fact that the non-linear drop across the switch at high signal levels was the cause of much of the distortion led us to the circuit in Fig. 3. This has a much better distortion figure: −87 dB (0.0045%) at a supply voltage of ±5 V. When the supply voltage was increased to ±7.5 V, the distortion could no longer be measured accurately. This would mean that this circuit is suitable for even the most exacting audio requirements, were it not for the channel

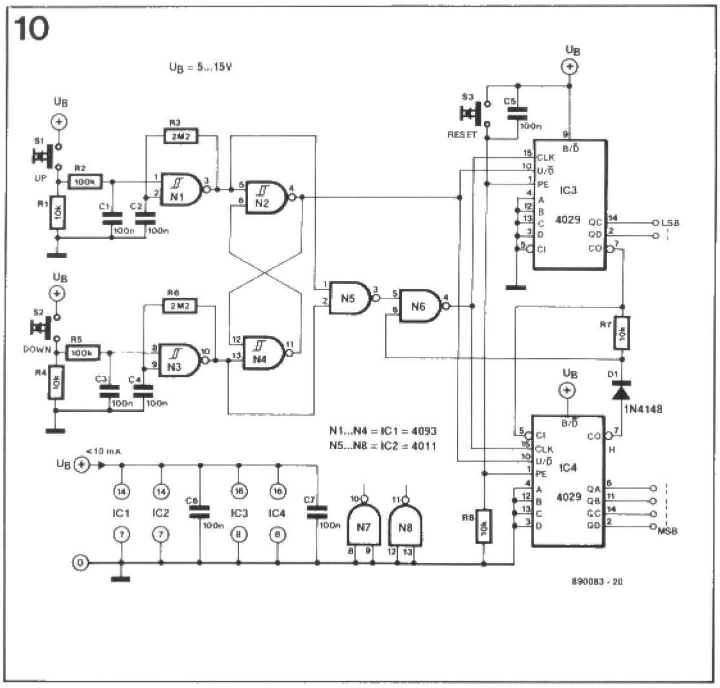


Fig. 10. Control circuit for Fig. 8.

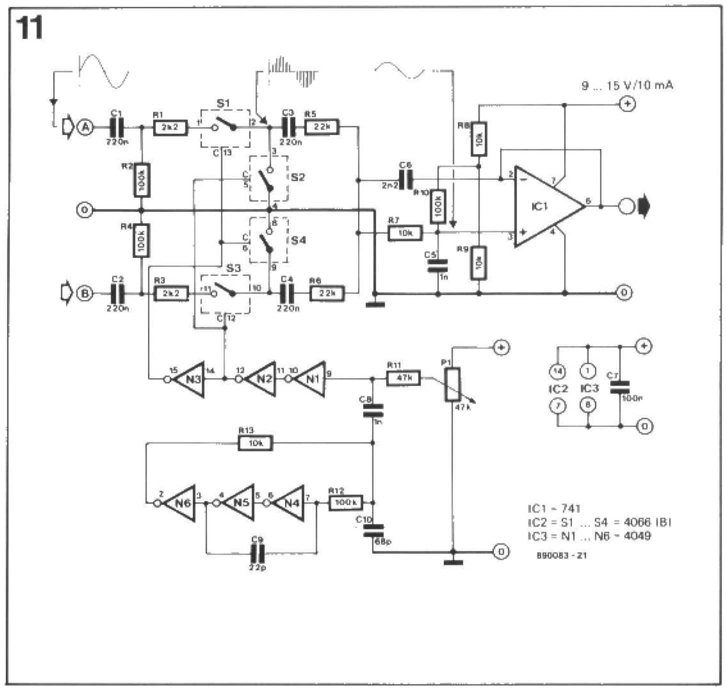


Fig. 11. This circuit could be considered a pulse-duration modulation mixer with CMOS switches.

Switches [1]:0,5 MHz [2]:1,25 MHz [3]:900 kHz [4]:1 MHz [5]:1 kHz [6]:100 kHz

Type	Maker	Functions	Ron [Ω]	Δ Ron [%]	RonMatch[%]	Uass [V]	ISOoff [dB]	CT [dB]	Ton [ns]	Toff [ns]	Remarks
SW01/02	PMI	4 × off/4 × on	85	7	4	26	58[1]	70	300	200	–
SW05		2 × off	45	5	5	26	62[1]	76	325	210	–
SW06		4 × on	60	5	5	26	58[1]	70	340	200	Disable input
SW201/202		4 × off/4 × on	60	5	5	26	58[1]	70	340	200	–
SW7510/11		4 × on/4 × off	60	15	1,5	26	66[1]	70	350	260	–
MM74HC4016	NS	4 × on	20	~50	25	15	44[4]	50	10	30	upgraded 4016
MM74HC4066		4 × off	60	~65	25	15	44[4]	50[4]	13	38	upgraded 4066
CD4066		4 × off	80	~40	6	15	50[2]	50[3]	50	50	–
AD7510/11	AD	4 × on/4 × off	75	20	1	24	*	*	180/350	350/180	
AD7512		2 × change-over	75	20	1	24	*	*	300	300	
AD7590/91/9201		4 × on/4 × off/ 2 × change-over	60	15	3	20	85[5]	56	240/400/350	400/250/350	with input Latches
DG300-303		see Fig. 12							150	130	
DG304-307	MAXIM		30	< 20	*	30	62[1]	74[1]	110	70	CMOS compatible
DG381-390									150	300	
IH5040-45			75	*	3	30	54	54	400	200	
IH5048-51			40	*	< 20	28	54	54	400	200	
IH5140-45			50	*	6	30	54	54	100 . . . 200	125 . . . 75	

Table 1. Essential data of some popular and interesting CMOS switches.

separation and crosstalk (−84 dB at 1 kHz; −60 dB at 20 kHz). Although the measured figures would be satisfactory for mass-produced equipment, they are not for good-quality apparatus. Typical characteristics of these parameters are given in Fig. 4 and Fig. 5. It may also be considered a drawback of the circuit that the opamp inverts the signal.

A further improvement of the circuit is shown in Fig. 6. This has an additional CMOS switch that short-circuits the signal when the switch in the signal path is open. The control signals for the two switches must therefore be in antiphase. The circuit shows an improvement in crosstalk and channel separation to −84 dB at 20 kHz. At this frequency the layout of the PCB makes a greater contribution to the distortion, as we have found many times in the design of audio equipment.

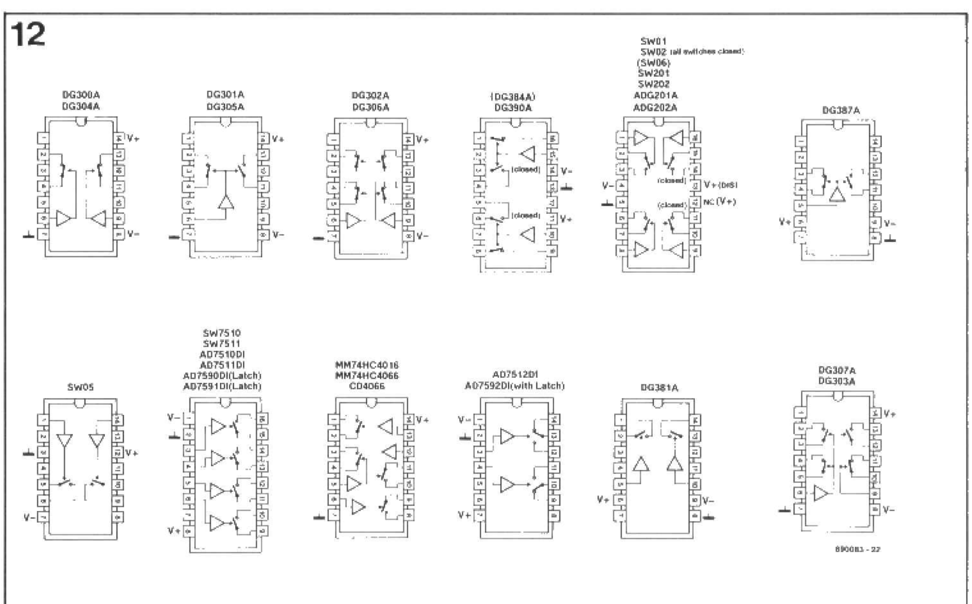


Fig. 12. Pin-out diagrams of the CMOS switches in Table 1.

Multiplexers

Type	Maker	Functions	Ron [Ω]	Δ Ron [%]	Ron Match [%]	Uass [V]	ISOoff [dB]	CT [dB]	Tus [μs]	Remarks
MUX08	PMI	1 × 1 off 8	220	1	7	25,4	60	70	1,8	–
MYX24		2 × 1 off 4	220	1	7	25,4	66	76	1,8	–
MUX16		1 × 1 off 16	290	1,5	7	26	66	75	1,7	–
MUX28		2 × 1 off 8	290	1,5	7	26	66	75	1,7	–
MUX88		1 × 1 off 8	220	1,5	12	36	88	98	1,8	3-bit binary; enable input
MM74HC4051	National	1 × 1 off 8								*: no data
MM74HC4052	Semiconductor	2 × 1 off 4	112	~ 60	~ 15	11	*	*	*	inhibit input
MM74HC4053		3 × 1 off 2								bi-directional
AD7501	Analog Devices	1 × 1 off 8								enable input
AD7502		2 × 1 off 4	170	20	4	25	*	*	1,6	enable input
AD7503		1 × 1 off 8								enable; no data
AD7506		1 × 1 off 16	300	15	4	25	70	*	1,6	enable input
AD7507		2 × 1 off 8								*: no data
IH6108	Intersil	1 × 1 off 8	300	20	*	28	60	*	1,2	enable input
IH6208		2 × 1 off 4								
DG508A	MAXIM	1 × 1 off 8	130	> 24	6	30	68	*	0,8	fast, bi-directional
DG509A		2 × 1 off 4								

Table 2. Essential data of some popular multiplexers.

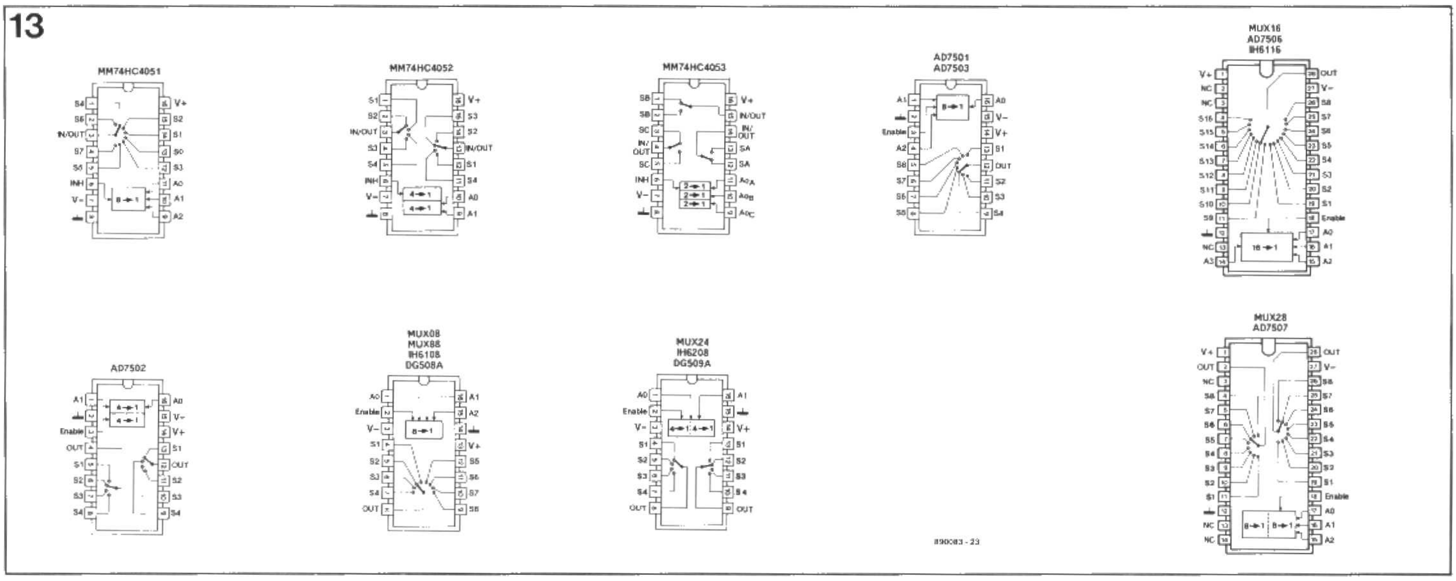


Fig. 13. Pin-out diagrams of the multiplexers in Table 2.

In the choice of a type of switch, quality, available space on the PCB, and price, play a role. If quality is deemed the most important factor, it is best to use single-switch ICs. On the other hand, if price is important, there are analogue multiplexers that contain a number of switches in one housing (just like stepping switches). These ICs save money and space. However, as will be seen from Table 2, a number of parameters of these devices are considerably worse than those of single-switch devices.

CMOS preamplifier

The circuit of Fig. 6 may be used to form an important part of a complete preamplifier, a basic design of which is shown in Fig. 7. The source selector may be a 2×1-form-8 multiplexer. The volume control may consist of two 1-from-8 multiplexers as shown in Fig. 8, or of single CMOS switches as shown in Fig. 9. If auxiliary functions, for instance, bass lift or stand by, are required, they may be realized with the aid of single CMOS switches.

The control logic is also fairly simple to design as shown in Fig. 10. This circuit is based on two start-stop oscillators, N_1 and N_3 respectively. NAND gates N_5 and N_6 generate the appropriate signal for 6-bit counter IC_3-IC_4. At the same time, the state of monostable N_2-N_4 determines whether IC_3 will count up or down. The outputs of the counter may be connected direct to the volume control in Fig. 8. Make sure that the level of the control signal to the logic circuits and of that to the CMOS switches are the same.

Volume control by signal ratio

An interesting application of fast CMOS switches is shown in Fig 11. The four switches are clocked by astable multivibrator N_1-N_6 at a frequency of 100-150 kHz (sampling theory holds that the clock frequency must be at least twice as high as the highest audio frequency).

Switches S_1 (S_4) and S_2 (S_3) are provided with control voltages that are in antiphase and are, therefore, never open or closed at the same time. The duty cycle is determined by the setting of P_1.

The 'lumps' of audio signal at the output of the switches are fed to IC_1. This opamp serves as a low-pass filter — (for removing the clock signal); as an integrator (for synthesizing the lumps of audio signal; and as an impedance converter.

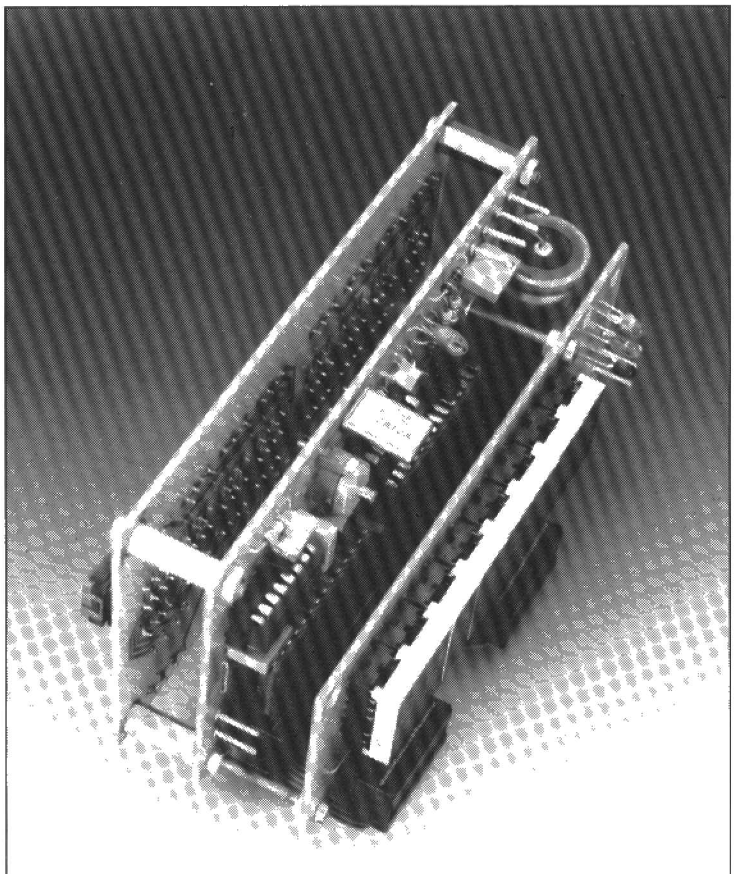
The circuit as shown receives two audio signals whose attenuation is inversely proportional to their loudness: the louder channel A, the softer channel B. Many variations may be applied to the circuit without affecting the original audio signal: one channel may be omitted; P_1 may be replaced by the circuits in Fig. 8 and Fig. 10; and others that we will leave to the reader's ingenuity.

8-DIGIT FREQUENCY METER

by T. Giffard

A state-of-the-art frequency meter module is presented that has an 8-digit, 7-segment LED indication, a resolution of 10 Hz, and accepts input frequencies of up to 3.5 MHz. Its presetting facility makes this simple-to-build module ideal for incorporation in a radio receiver.

The module is based on two ICM7217IPL CMOS presettable up/down counters. Two of these chips are cascaded to obtain an 8-digit read-out on common-anode 7-segment LED displays.



The counter's presetting facility makes it eminently suitable for use as a frequency read-out in receivers, since the intermediate frequency (e.g., 455 kHz or 9 MHz, can be programmed as an offset. In this manner, the output frequency of the local oscillator (L.O.) may be measured by the counter module, when driven by a suitable prescaler. Depending on whether the L.O. frequency is lower or higher than the received frequency, the IF offset is divided by the prescale ratio and then programmed as a preset value, which is automatically added to, or subtracted from, the module's input frequency to ensure that the *received frequency* is shown on the display.

An example might help to illustrate the above procedure. A super-heterodyne VHF FM broadcast receiver has an intermediate frequency of 10.7 MHz. The L.O. frequency is higher than the received frequency. Assuming that the receiver is tuned to a station at 100.0 MHz, the L.O. generates 110.7 MHz. This signal is applied to a divide-by-100 prescaler, which drives the frequency meter module. To ensure that the display reads 100 MHz, the counter must be programmed for an IF offset of 10.7 MHz/100=107 kHz. Since the counter will normally count up, it must be set to a negative offset, the *one's-complement* of this frequency, which is simple to calculate as

10 000 000 – 0 107 000 = 09 893 000.
shift right (10 Hz); MSD borrow;
preset = 99 989 300

The counter module has an up/down input and a separate, but optional, circuit for programming the offset. Resolution and gating times are simple to change, if desired. The maximum input frequency of the counter module is about 3.5 MHz at a sensitivity of 60 mV_{rms}.

The counter chip

The ICM7217IPI is a CMOS decade counter in a 28-pin plastic enclosure, intended for being programmed with the aid of switches or fixed logic configurations, and driving common-anode displays. The device from GE-Intersil (second source: Maxim) is one of a family of single-chip 4-bit programmable up/down counters with an on-chip multiplex scan oscillator for simple driving of 7-segment LED displays.

The internal structure of the ICM7217 is given in Fig. 1. Three main outputs are provided: CARRY/BORROW for cascading with further 4-bit counters, ZERO which indicates when counter state zero (0000) is

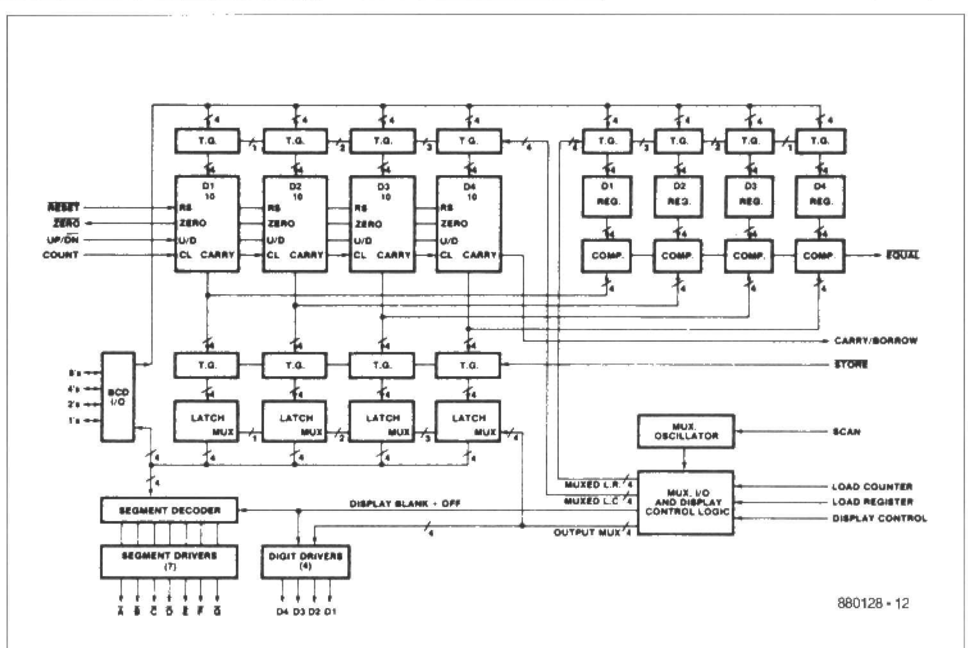


Fig. 1. Block diagram of the ICM7117 (*courtesy GE-Intersil*).

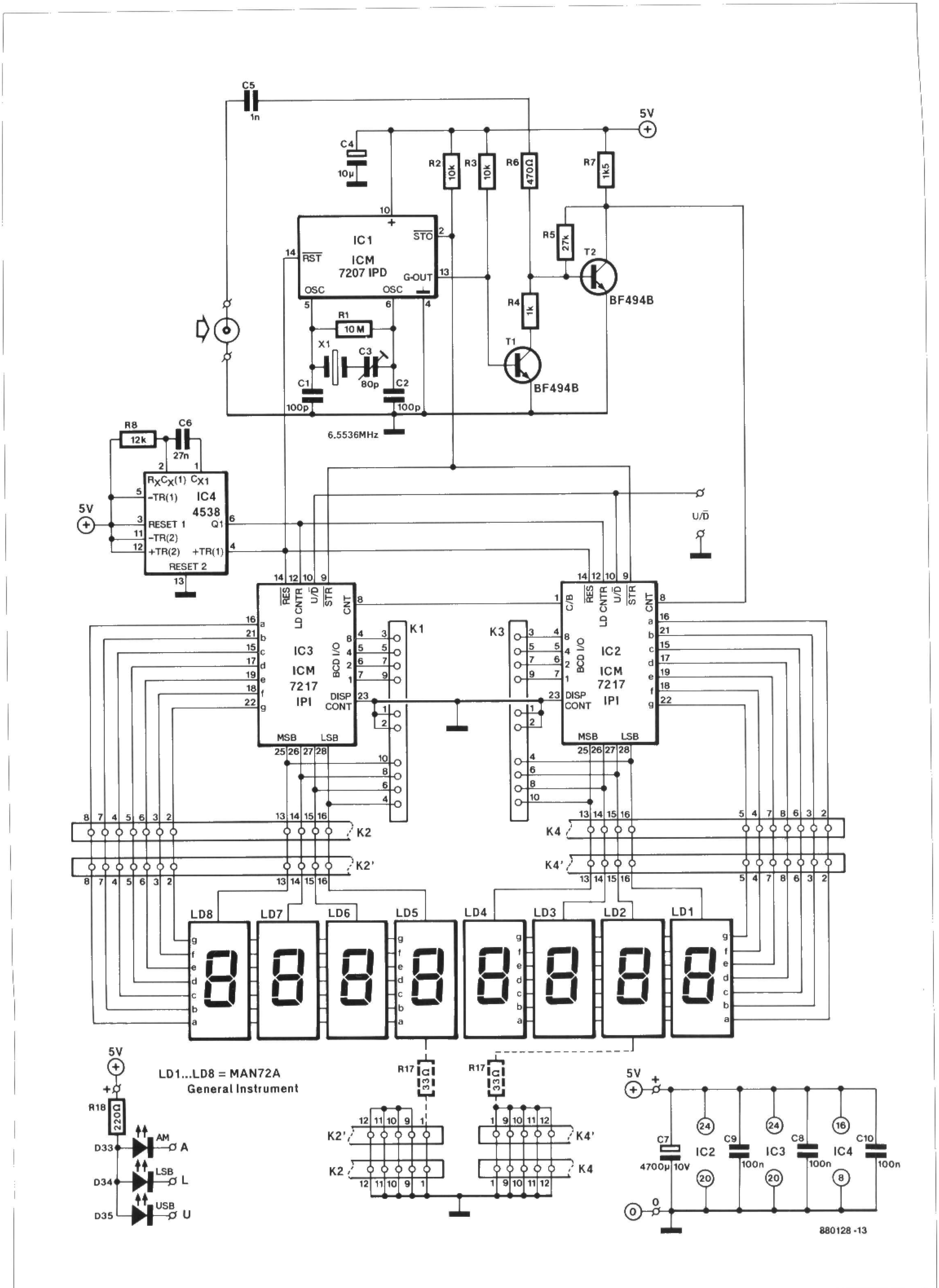


Fig. 2. Circuit diagram of the presettable 8-digit counter module with up/down input and LED read-out.

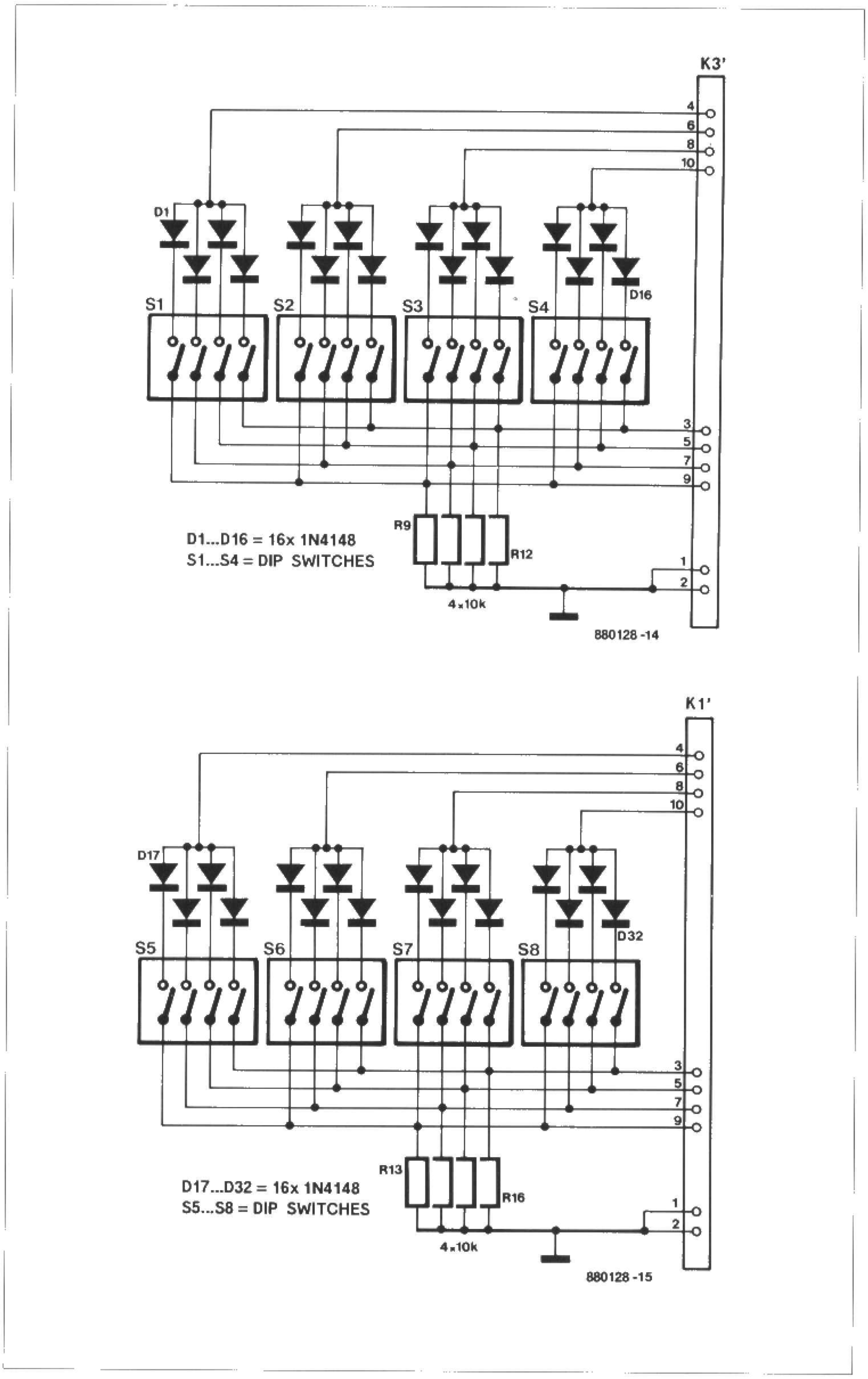


Fig.. 3. Circuit diagram of the optional preset unit.

reached, and EQUAL which indicates when the current counter state equals the value loaded into the internal register via the BCD I/O pins. The three outputs and the BCD port are TTL-compatible and internally multiplexed. Output CARRY/BORROW goes high when the counter is clocked from 9999 to 0000 when counting up (input U/D logic high), or from 0000 to 9999 when counting down (input U/D logic low). The Schmitt-trigger at the COUNT input provides hysteresis to prevent double clocking on slow rising edges.

The counter contents are transferred to the multiplexed 7-segment and BCD outputs when input STORE is pulled low. A low level at the RESET input causes the counter to be asynchronously reset to 0000.

As already noted, the BCD port can function as an input or an output. These functions are selected with the logic levels applied to the three-level LOAD COUNTER (LD) and LOAD REGISTER (LR) inputs. When both are open, the BCD port provides the multiplexed BCD display selection signals, scanning from MSD (most-significant display) to LSD (least-significant display). When either LR or LC is taken high, the BCD port is turned into a 4-bit input for loading the counter (LC) or register (LR) data. Since the ICM7217IPI is designed to drive common-anode displays, the levels applied to, or provided by, the BCD port are 'high true'.

When input LR is made low, the BCD I/O lines are switched to the high-impedance state, and the digit and segment drivers are turned off. The counting operation continues, however, and the remaining input and output functions operate normally. The displays are normally switched off with the aid of input LR to reduce power consumption during standby conditions.

The on-board multiplex scan oscillator controls the internal timing of the ICM7217. The nominal oscillation frequency of 2.5 kHz may be reduced by connecting a capacitor between input SCAN and the positive supply line. The oscillator output signal has a relatively low duty factor to delay the digit driver outputs and thus prevent 'ghosting' effects on the displays.

The digit and segment drivers on board the ICM7217 are capable of directly driving common-anode 7-segment LED displays at a peak segment current of 40 mA. At a duty factor of 0.25, this corresponds to 10 mA per segment.

Finally, the DISPLAY CONTROL input recognizes 3 logic levels. When it is logic high, the display segments are inhibited. When it is logic low, the leading zero blanking feature is turned off. Displays on with leading zero suppression is achieved by leaving the input open.

Practical circuit

As shown in the circuit diagram of Fig. 2, a pair of ICM7217IPIs is used in conjunction with a central timing generator type ICM7207IPD (IC_1). This chip controls the gating of the input signal with the aid of an external quartz crystal, X_1, inverter T_1 and input amplifier T_2. In addition, the ICM7207IPD provides the STORE and RESET signal for the counter chips, IC_2 and IC_3. Although the STORE output of the ICM7207IPD is of the open-drain type, and the associated inputs of the ICM7217s have 75 µA pull-up resistors, an external pull-up resistor R_2, is fitted to ensure immunity to noise. The U/D and RESET inputs also have internal pull-up resistors, and may, therefore, be left open for normal operation as an up-counter. The block diagram of the ICM7207 is given in Fig. 4.

Monostable IC_4 enables the counter to load the preset word. The LOAD COUNTER pulse is delayed with respect to the RESET pulse because the counter can only be preset with data other than 0000 when RESET is inactive.

The preset frequency is set with two blocks of 4-way DIP switch blocks. The circuit diagrams of these (optional) units are given in Fig. 3. BCD thumbwheel switches may be used as a more ergonomical alternative to the DIP switches. Alternatively, wire links may be used if the counter works with one, fixed, preset frequency.

The BCD port lines and the scanning

digit selection signals are available on K_1 and K_3 for connecting to the preset unit.

A few suggestions are given for those who want to experiment with the circuit. The duration of the count window may be reduced from 100 ms to 10 ms by tying pin 11 of the ICM7107IPD (RANGE CONTROL) to the positive supply line. This modification results in a corresponding reduction of the counter's resolution, however: with pin 11 at +5 V, this is 100 Hz instead of 10 Hz. In both cases, a good-quality 6.5536 MHz quartz crystal is required: for optimum stability of the read-out, a type with 10 ppm tolerance or better is recommended (most inexpensive computer crystals do not meet this specification).

For high-resolution applications, the duration of the count window may be increased by a factor 10 (100 ms or 1 s) by using a ICM7207A in combination with a 5.24288 MHz quartz crystal. Unfortunately, this is not a standard frequency, so that this crystal will have to be made to order.

Pin 23 of both counter chips is connected to ground, so that leading-zero suppression is not used. As already discussed, this feature may be useful in a number of applications. Where it is required, pin 23 of IC_3 may be left open to achieve leading-zero suppression on the most-significant display group. Leading-zero suppression of the full 8-digit display may be realized by driving the DISPLAY CONTROL input pin of the LS group driver, IC_2, with the collector signal of a n-p-n transistor whose base is driven by the ZERO output of the MS group driver, IC_3. In a number of cases, it may be possible to omit the two MS displays altogether.

Resistor R_{17} is only required when the module is used without a prescaler. Depending on whether a MHz or kHz indication is required, the resistor lights the decimal point on LD_5 (MHz read-out) or LD_2 (kHz read-out).

Three receiver mode indicators, D_{33}, D_{34} and D_{35}, are provided on the display board. The LEDs may be controlled from the mode selection switch in the receiver.

Table 1

Switch block	resolution (Hz)	multiplier
S1	1	× 10
	2	
	4	
	8	
S2	1	× 10^2
	2	
	4	
	8	
S3	1	× 10^3
	2	
	4	
	8	
S4	1	× 10^4
	2	
	4	
	8	
S5	1	× 10^5
	2	
	4	
	8	
S6	1	× 10^6
	2	
	4	
	8	
S7	1	× 10^7
	2	
	4	
	8	
S8	1	× 10^8
	2	
	4	
	8	

Example 1:
IF = 455 kHz; $f_{LO} > f_i$; no prescaler;
counter mode: UP.
Preset = 99954500
Switches set to 'on':
S3(1) and (4); S4(4); S5(1) and (4); S6(1) and (8); S7(1) and (8); S8(1) and (8).

For $f_{LO} < f_i$:
preset = 45500
DIP switches set to 'on':
S3(1) and (4); S4(1) and (4); S5(4)

Example 2:
IF = 9 MHz; $f_{LO} > f_i$; prescaler ÷10;
counter mode: UP.

Preset = 99100000
DIP switches set to 'on':
S6(1); S7(1) and (8); S8(1) and (8)

For $f_{LO} < f_i$:
preset = 900000
DIP switches set to 'on':
S6(1) and (8)

Example 3:
IF = 10.7 MHz; $f_{LO} > f_i$; prescaler ÷100;
counter mode UP.
Preset = 99989300
DIP switches set to 'on':
S3(1) and (2); S4(1) and (8); S5(8); S6(1) and (8); S7(1) and (8), S8(1) and (8).

For $f_{LO} < f_i$:
DIP switches set to 'on':
S3(1) and (2) and (4); S5(1)

Three boards: a compact frequency read-out

The lay-out of the printed-circuit board for the universal counter is given in Fig. 5. The PCB is cut into three to separate the preset unit (at the top), the main counter board (at the centre), and the read-out section (at the bottom). The receiver mode indication board forms a separate unit, which need, however, not be cut from the display board.

Populating the boards is straightforward and requires hardly any comment. It is strongly recommended to use sockets for all integrated circuits, displays and DIP switches. K_2' and K_4' on the display board, and K_2 and K_4 on the main counter board, are 16-way IC sockets with turned pins. These receive 16-way IDC pin-headers fitted at the ends of an approximately 5 cm long flat-ribbon cable. The 10-way connections between the main counter board and the read-out are made in 10-way flat-ribbon cables.

Pin-headers K_1 and K_3 on the main counter board are fitted *at the component side*, and K_1' and K_3' on the preset board *at the track side*. The pin-headers are connected with IDC sockets pressed on to the ends of an approximately 5 cm long flat-ribbon cable.

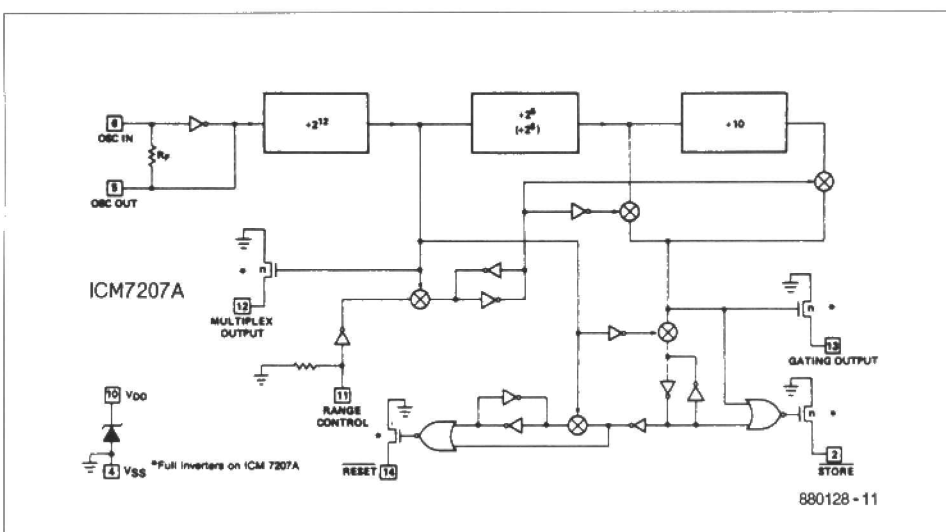


Fig. 4. Internal structure of the ICM7207A timing generator (*courtesy GE-Intersil*).

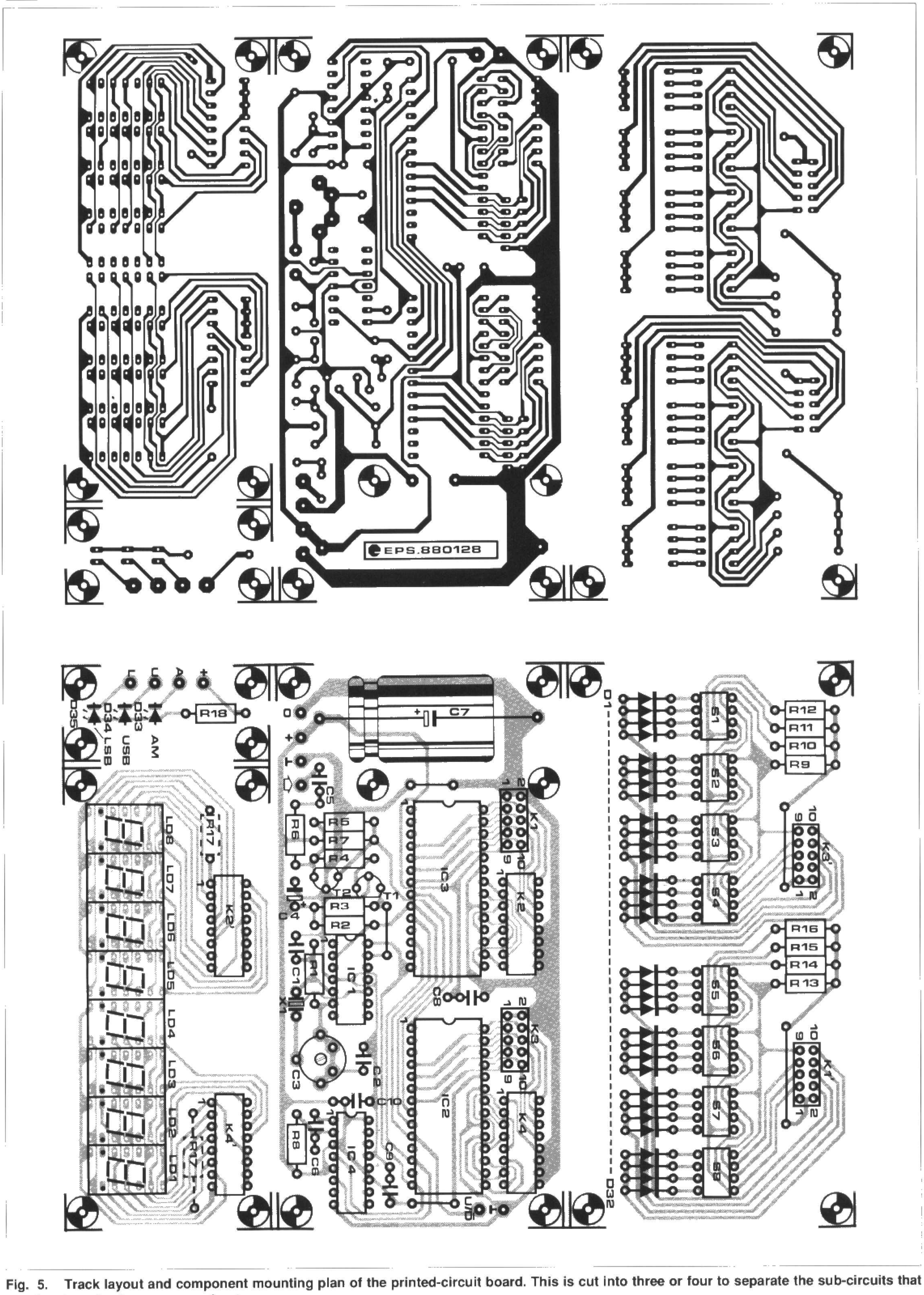


Fig. 5. Track layout and component mounting plan of the printed-circuit board. This is cut into three or four to separate the sub-circuits that together form the frequency read-out.

Parts list

Resistors (±5%):
R_1=10M
R_2;R_3;R_9–R_{16} incl.=10K
R_4=1K0
R_5=27K
R_6=470R
R_7=1K5
R_8=12K
R_{17}=33R (see text)
R_{18}=220R

Capacitors:
C_1;C_2=100p
C_3=80p trimmer
C_4=10µ; 25 V; tantalum
C_5=1n0
C_6=27n
C_7=4700µ; 10 V
C_8;C_9;C_{10}=100n

Semiconductors:
D_1–D_{32} incl.=1N4148
D_{33};D_{34};D_{35}= LED
IC_1=ICM7207IPD (GE-Intersil or Maxim)
IC_2;IC_3=ICM7217IPI or ICM7217IPJ (GE-Intersil or Maxim)
IC_4=4538
LD_1–LD_8 incl.=MAN72A (General Instrument Optoelectronics)
T_1;T_2=BF494B

Miscellaneous:
K_2;K_2';K_4;K_4'= 16-way DIL socket with mating IDC plug.
K_1;K_1';K_3;K_3'= 10-way pin header with mating IDC socket.
S_1–S_8 incl.= 4-way DIL switch block.
X_1= 6.5536 MHz quartz crystal.
PCB Type 880128 (see Readers Services page).

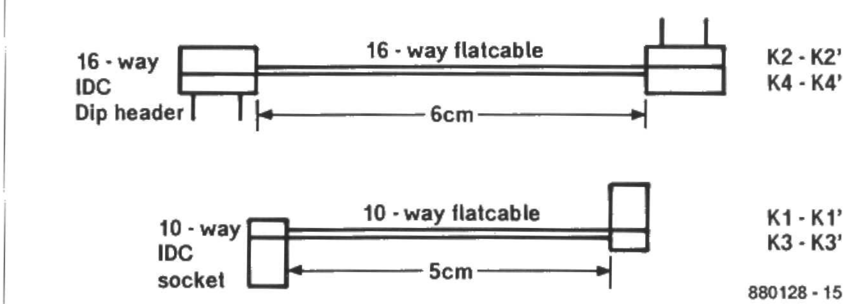


Fig. 6. Construction of the four flat-ribbon cables that interconnect the sandwiched boards.

The construction of the flat-ribbon cables that interconnect the sub-modules is illustrated in Fig. 6. Contrary to what some retailers of specialist tools would have you believe, IDC (*insulation displacement*) connectors are simple to fit on to flat-ribbon cable with the aid of a carefully operated vice, or even a small hammer and two pieces of wood. Insert the cable between the socket or plug and the associated plastic cap, and align the individual wires with the clip-type connectors. Then close the connector by carefully pressing the cap on to body of the connector. Alternatively, carefully tap the cap in place with the aid of a small hammer. Check the continuity at all pins.

The completed sub-assemblies are then ready for mounting together in a sandwich construction. The read-out board is mounted on top of the main counter board with the aid of three 25 mm long spacers or lengths of M3 threading. Make sure that the soldering connections of the receiver mode LEDs, and those for the nearby terminal posts, do not touch the body of the large electrolytic capacitor, C_7, underneath. The preset board is fitted back-to-back below the main counter board with the aid of 20 mm long PCB spacers with internal threading. The completed three-board assembly is shown in the introductory photograph of this article.

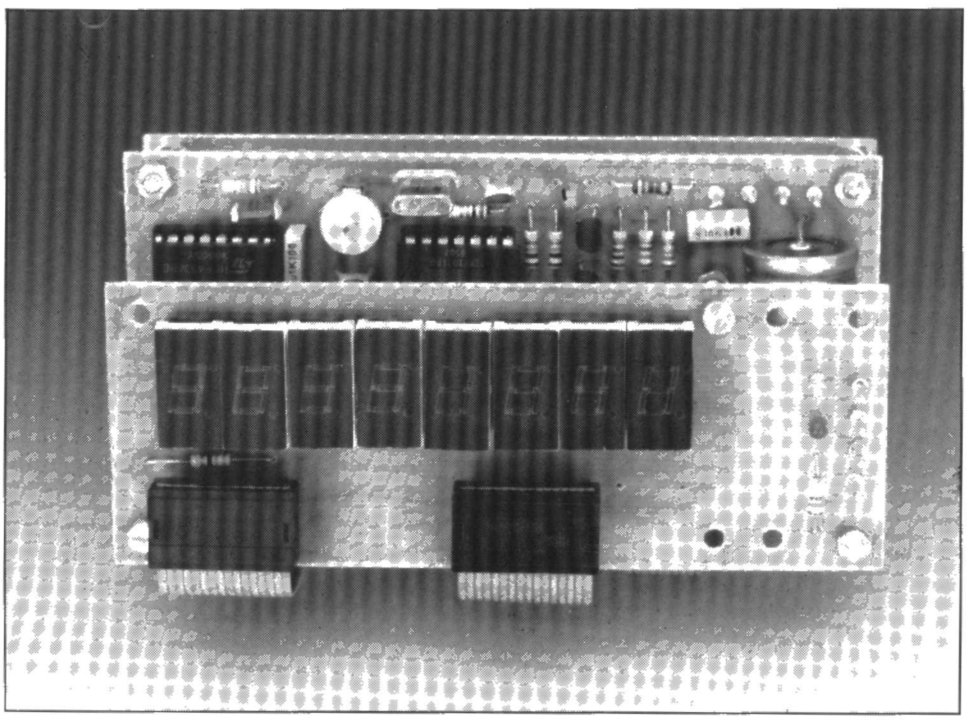
The unit may be installed in a receiver and connected to a regulated and well-decoupled 5 V power supply. In some cases, it may be necessary to screen the module to prevent interference in the receiver. The readability of the displays may be improved by fitting them behind a red bezel.

Calibration is simple if a frequency meter is available: adjust trimmer C_3 for 6.5536 MHz measured at pin 5 of the ICM7207. Alternatively, tune the receiver for zero-beat against a frequency reference station, and adjust the trimmer until the correct received frequency is displayed.

Sensitivity of the prototype was 35 mV_{rms} over 200 kHz to 1 MHz, and 60 mV_{rms} at an input frequency of 3 MHz. Average current consumption with eight displays on (indication: 8×'8'), but the receiver mode LEDs off, was measured at approximately 450 mA.

Offset programming

Assuming that the counter operates in the UP mode, and that the local oscillator frequency is higher than the received frequency, the required preset value is first converted to its 8-digit one's complement. Next, the corresponding DIP switches are set until the preset appears on the displays. Examples for 455 KHz, 900 kHz (9 MHz with ÷10 prescaler) and 107 kHz (10.7 MHz with ÷100 prescaler) are given in Table 1. Always remember that the counter can not handle input frequencies higher than 3.5 MHz, so that the effectively programmed offset is the IF frequency divided by the prescale factor. For most SW and general coverage receivers, a ÷10 prescaler is suitable; for VHF receivers a ÷100 prescaler.



PRACTICAL FILTER DESIGN – PART 6

by H. Baggott

In this sixth part in the series we start our discourse of the tables and characteristics of filters and as first we deal with those pertaining to the Butterworth type because that is the best known and probably also the most often used kind of filter.

The Butterworth filter owes its popularity to a combination of flat amplitude response in the pass band and reasonable roll-off. A drawback is its non-linear phase characteristic.

The roll-off is fairly precisely 6*n* dB per octave, where *n* is the order of the filter.

The Butterworth filter may be considered a compromise between the Bessel network (moderate roll-off but linear phase response) and the Chebishev filter (steep roll-off, poor phase response and ripple in the pass band). For applications that require a flat pass band and steep roll-off, the Butterworth filter is undoubtedly the best choice.

Table 1 gives the pole locations of Butterworth filters of the second to the tenth order. These data enable the ready computation of filters with the aid of formulas given in earlier parts in this series.

Butterworth tables

The dimensioning of filters becomes much simpler with the aid of Tables 2 to 5,which give component values for passive and active filters of the second to the tenth order. The values given always refer to a filter with a cut-off frequency of 1 Hz.

Table 2 gives component values for a passive filter with identical source and output impedances. The component identications at the top of the table correspond to those in the diagrams above the table and those at the bottom of the table correspond to the diagrams below the table.

Table 3 gives the component values for a passive filter with negligible source impedance.

Tables 4 and 5 give the component values for active filters with a single feedback path. Table 4 deals with second- and third-order sections. If, for instance, you want to design a seventh-order filter, you take two second-order and one third-order section and connect them in tandem.

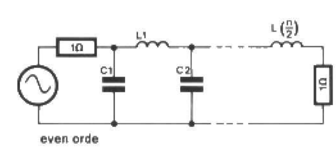
It is also possible, as we have seen in Part 3, to use only second-order sections and, in the case of odd-order filters, add a passive *RC* network. The data for this are shown in Table 5. This table is given merely to illustrate the alternative way. Since in the majority of cases it is simpler to work with Table 4, Table 5 will not be given for the other filter types in future parts in this series.

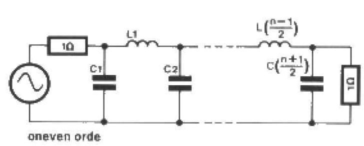
Butterworth characteristics

For clarity's sake, the characteristics given in this article deviate slightly from those given as examples in Part 2. For each type of filter we will give three series of characteristics, showing respectively: the gain vs frequency response—Fig.32; the delay vs frequency response—Fig. 33; and the step vs time response—Fig. 34. The phase response is not given because this would not divulge all that much on a logarithmic scale. In any case, the phase linearity is easily deduced from Fig. 33, since linearity corresponds to a constant delay time at

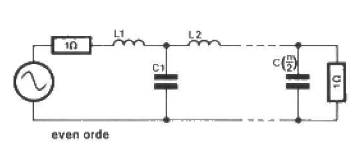
n	real part $-\alpha$	imaginary part $\pm\beta$
2	0.70711	0.70711
3	0.5 1	0.86603
4	0.38268 0.92388	0.92388 0.38268
5	0.30902 0.80902 1	0.95106 0.58779
6	0.25882 0.70711 0.96593	0.96593 0.70711 0.25882
7	0.22252 0.62349 0.90097 1	0.97493 0.78183 0.43388
8	0.19509 0.55557 0.83147 0.98079	0.98079 0.83147 0.55557 0.19509
9	0.17365 0.5 0.76604 0.93969 1	0.98481 0.86603 0.64279 0.34202
10	0.15643 0.45399 0.70711 0.89101 0.98769	0.98769 0.89101 0.70711 0.45399 0.15643

Table 1. Pole locations of Butterworth filters.





n	C1	L1	C2	L2	C3	L3	C4	L4	C5	L5
2	0.2251	0.2251								
3	0.1592	0.3183	0.1592							
4	0.1218	0.2941	0.2941	0.1218						
5	0.09836	0.2575	0.3183	0.2875	0.09836					
6	0.08238	0.2251	0.3075	0.3075	0.2251	0.08283				
7	0.07083	0.1985	0.2868	0.3183	0.2868	0.1985	0.07083			
8	0.0621	0.1768	0.2647	0.3122	0.3122	0.2647	0.1768	0.0621		
9	0.05527	0.1592	0.2438	0.2991	0.3183	0.2991	0.2438	0.1592	0.05527	
10	0.04979	0.1445	0.2251	0.2836	0.3144	0.3144	0.2836	0.2251	0.1445	0.04979
	L1	C1	L2	C2	L3	C3	L4	C4	L5	C5



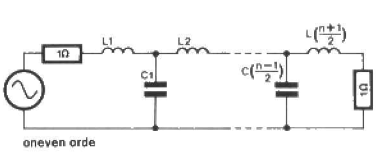
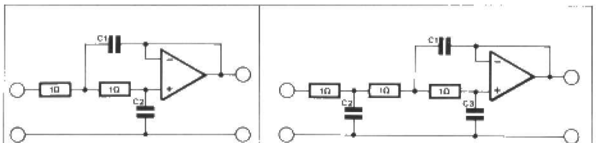


Table. 2. Normalized component values for passive low-pass filters with identical input and output impedances.

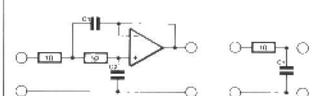
n	L1	C1	L2	C2	L3	C3	L4	C4	L5	C5
2	0.2251	0.1126								
3	0.2387	0.2122	0.07985							
4	0.2436	0.251	0.1732	0.06091						
5	0.2459	0.2697	0.22	0.1423	0.04918					
6	0.2472	0.28	0.2472	0.1912	0.1206	0.04119				
7	0.2479	0.2836	0.264	0.2224	0.1679	0.1044	0.03541			
8	0.2484	0.2904	0.2751	0.2432	0.2003	0.1491	0.09193	0.03105		
9	0.2487	0.2932	0.2829	0.2579	0.2234	0.1816	0.1339	0.08204	0.02763	
10	0.249	0.2953	0.2884	0.2685	0.2414	0.2056	0.1656	0.1214	0.07407	0.02489

Table 3. Normalized component values for passive low-pass sections with negligible source impedance.



n	C1	C2	C1	C2	C3
2	0.2251	0.1125			
3			0.5644	0.2215	0.03221
4	0.4159 0.1723	0.06091 0.147			
5	0.515	0.04918	0.279	0.2155	0.06707
6	0.6149 0.2251 0.1648	0.04119 0.1125 0.1537			
7	0.7152 0.2553	0.03542 0.09923	0.2437	0.2126	0.07775
8	0.8158 0.2865 0.1914 0.1623	0.03105 0.08842 0.1323 0.1561			
9	0.9165 0.3183 0.2078	0.02764 0.07958 0.1219	0.2316	0.2112	0.08228
10	1.0174 0.3506 0.2251 0.1786 0.1611	0.0249 0.07225 0.1125 0.1418 0.1572			

Table. 4. Normalized component values for active filters with single feedback path.



n	C1	C2	C1
2	0.2251	0.1125	
3	0.3183	0.07958	0.1592
4	0.4159 0.1723	0.06091 0.147	
5	0.515 0.1967	0.04918 0.1288	0.1592
6	0.6149 0.2251 0.1648	0.04119 0.1125 0.1537	
7	0.7152 0.2553 0.1766	0.03542 0.09923 0.1434	0.1592
8	0.8158 0.2865 0.1914 0.1623	0.03105 0.08842 0.1323 0.1561	
9	0.9165 0.3183 0.2078 0.1694	0.02764 0.07958 0.1219 0.1496	0.1592
10	1.0174 0.3506 0.2251 0.1786 0.1611	0.0249 0.07225 0.1125 0.1418 0.1	

Table 5. Normalized component values for filters with single feedback path.

all frequencies. Each of the figures gives the characteristics for a second-, fourth-, sixth-, eighth- and tenth-order section. Those for odd-order filters are assessed from intermediate values: this keeps the number of characteristics to a reasonable level to prevent loss of clarity.

Note that in Fig. 32 for a clear view of the behaviour of the filter just below the cut-off frequency, the scale of the *y*-axis to the left of 1 Hz has been expanded and is shown at the left of the drawing. The values of the gain at frequencies above 1 Hz are shown to the right of the drawing.

Two examples

We shall give a couple of worked out examples for each type of filter we deal with to give you the opportunity of learning to use the tables and characteristics quickly and properly.

Example 1.

Design a passive low-pass Butter-worth filter with a cut-off frequency of 1600 kHz and a source and output impedance of 50Ω. The attenuation at 3200 kHz must be at least 20 dB.

Solution

First we determine the value of the attenuation at each frequency relative to the normalized frequency of 1 Hz by dividing the reference frequency by the cut-off frequency:

3200 : 1600 = 2.

From Fig. 32 we determine which curve affords at least 20 dB attenuation at f=2 Hz. and this is found to be for a fourth-order filter the diagram of which is shown in Fig. 35a. Note that a third-order filter just would not do since it would give an attenuation of only 18 dB per octave.

It would also have been possible to deduce the filter from the diagram underneath Table 2. Study this carefully, because once you understand this, the purpose of Table 2 will be clear forever.

All that remains to be done now is to calculate the component values for the given input and output impedance and the cut-off frequency:

$$C' = C / (fR)$$

$$L' = L R / f$$

The calculations will be found to result in the component values given in the diagrams in Fig. 35b.

Similarly, the values for the components in Fig. 4a are found to be:

$$C_1 = 0.1218 / (1600000 \times 50) =$$

$$= 1.52 \times 10^{-9} = 1.52 \text{ nF}$$

$$L_1 = 0.2941\ (50 / 1600000) =$$

$$= 9.19 \times 10^{-6} = 9.19\ \mu\text{H}$$

Example 2.

Design an active fifth-order low-pass Butterworth filter with a cut-off frequency of

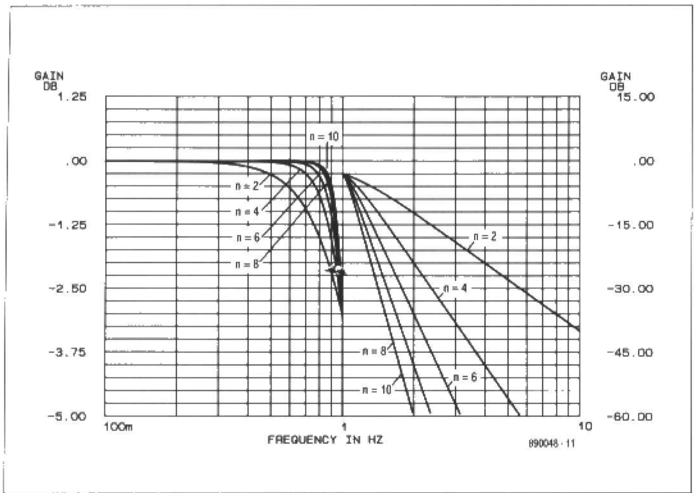


Fig. 32. Gain vs frequency characteristics of a Butterworth filter.

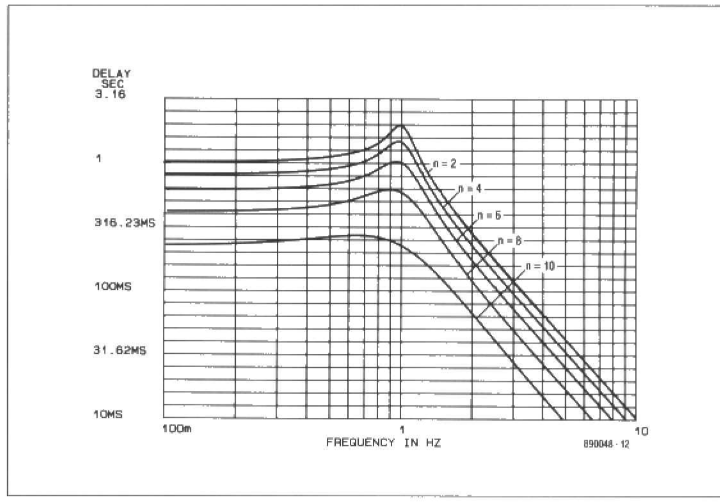


Fig. 33. Delay time vs frequency characteristics of a Butterworth filter.

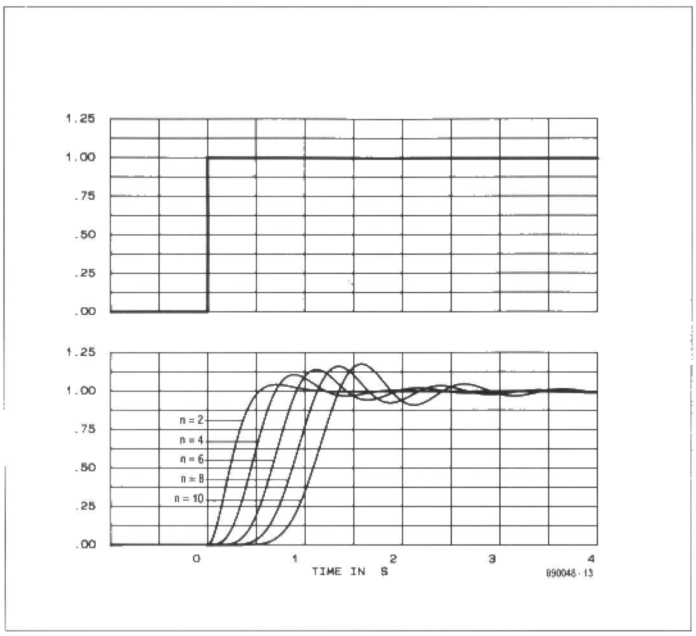


Fig. 34. Step response of a Butterworth filter

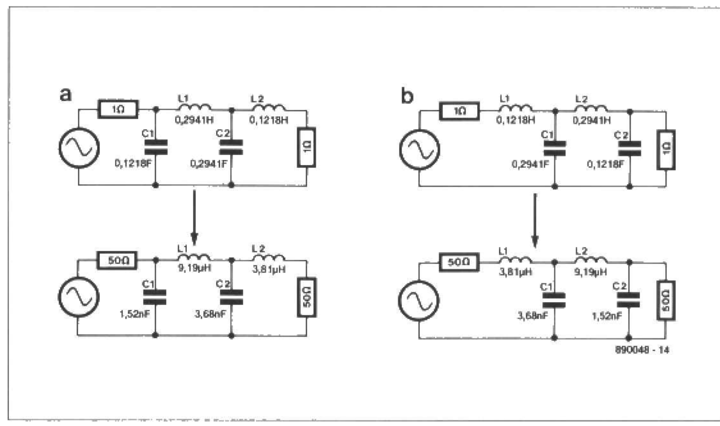
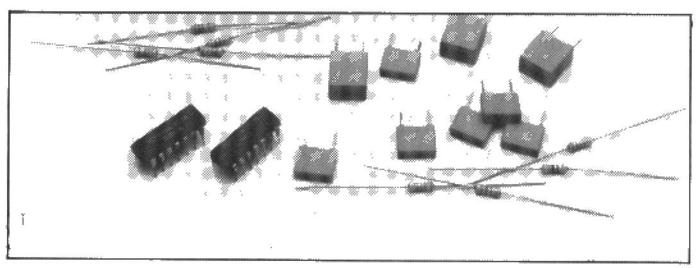


Fig. 35. Two examples of how to dimension a passive Butterworth filter.



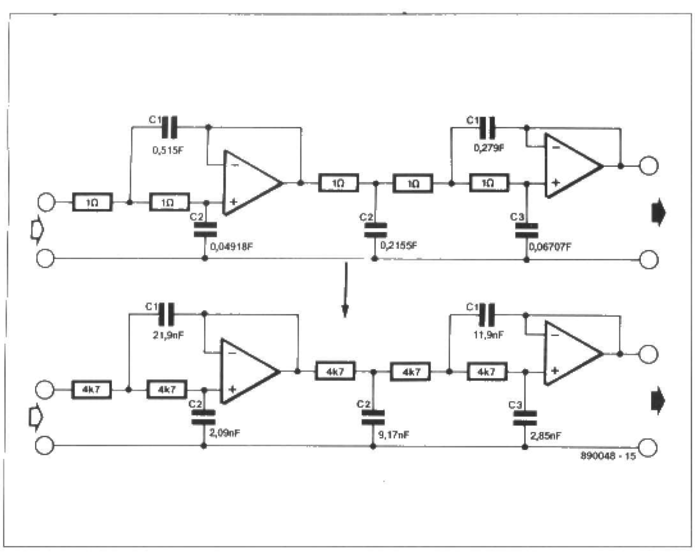


Fig. 36. Illustrating the computation of an active 5th-order filter.

5 kHz.

Solution.

This is designed fairly quickly. It is an odd-order filter, so we need a second-order section and a third-order section, as drawn above Table 4. The two sections are connected in tandem, after which the normalized component values read from the table are inserted.

Next, choose a value for the resistors (*R* in the formulas), say, 4.7 kΩ.

Then calculate with the aid of the formula given in the first example (for *C'*) the 'real' values of the components.

Again, two examples of the calculations:

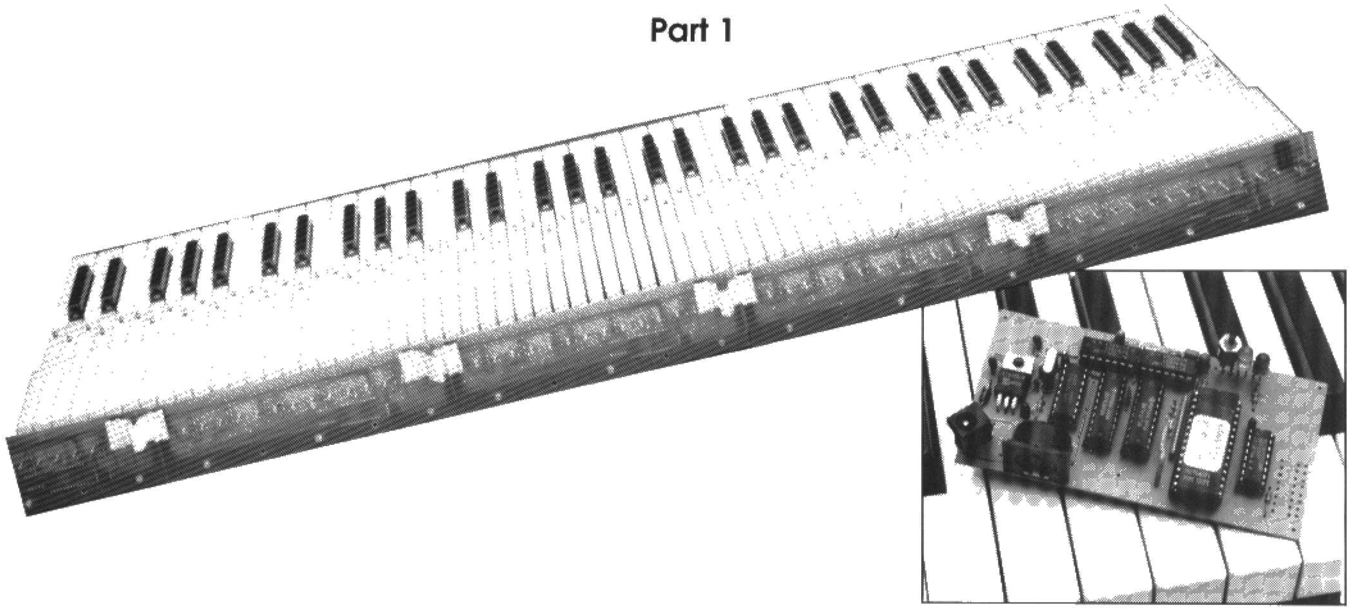
$$C_1 = 0.515 / (5000 \times 4700) = 21.9 \times 10^{-9} = 21.9 \text{ nF}$$

$$C_2 = 0.04918 / (5000 \times 4700 = 2.09 \times 10^{-9} = 2.09 \text{ nF}$$

This completes our discourse on Butterworth filters. Part 7 will deal with Bessel networks.

UNIVERSAL MIDI KEYBOARD INTERFACE

Part 1



by D. Doepfer

The feature *par excellence* of the MIDI-compatible keyboard controller described in this article is its ability to be used with practically any existing keyboard, whether salvaged from a discarded musical instrument, or still in function in a piano, organ, or non-MIDI synthesizer.

Soon after the publication of the *Portable MIDI keyboard* (Ref. 1), numerous readers asked us to give further details on the use of the Type E510 MIDI controller in conjunction with full-size keyboards of five and more octaves. This month we meet these requests with the description of a universal MIDI controller board, once again based on the E510, intended for use with many types of musical keyboard.

The maximum number of keys supported by the present design is no fewer than 96, covering 8 octaves. The controller provides the velocity parameter, and supports one-octave transposition as well as instantaneous split-point programming to achieve data distribution between MIDI channels 1 and 2, with any key on the keyboard. The printed-circuit boards have been designed such that they may be used in conjunction with a keyboard having wooden keys and spring- or gold-wire contacts (Kimber-Allen type). Any other type of key or contact is, however, also suitable.

A MIDI keyboard is classified as accessory equipment, not as an instrument, because it is not capable of producing musical sounds. As such, it is used for controlling MIDI synthesizers (*expanders*), or micro-processor based systems running special MIDI programs.

The application range of the present circuit is widened further by the fact that the key inputs are suitable for driving from almost anything that represents an electrical contact. We have, therefore, no reservations about calling the circuit *universally applicable.* To mention a few less usual, but technically interesting, applications: key signals generated by the player interrupting light-beams, or actuation by weight of touch-sensitive areas on a theatre or dance floor.

The velocity parameter is not always required for such applications, and is fairly simple to omit as will be shown later. Other ways of providing the key signals may come to your mind at this stage. At the end of the article, we describe an experimental percussion interface to rouse your interest in finding new applications for the MIDI controller.

We feel sure that the design will please many of our readers, who, nodoubt, will have their own follow-up suggestions for, say, a semitone transposition circuit, a sustain pedal, and typical MIDI functions such as program change, pitch bend and access to all 16 available channels. Let us know of such thoughts and ideas and we will respond appropriately.

This two-part article describes the operation, construction and use of the universal MIDI keyboard. Although space did not permit a reiteration of the introduction to the MIDI keyboard, a description of its principles and functions may be found in Ref. 1. This also discusses the way in which a MIDI keyboard controller circuit measures the time between the instant the pole of the key leaves its rest position and the instant it reaches the-

UNIVERSAL MIDI KEYBOARD

- universal polyphonic MIDI keyboard with a maximum of 8 octaves (96 keys)
- transmits velocity parameter
- 1 instantaneously programmable split-point (channels 1 and 2)
- ±1-octave transposition
- simple-to-build circuit
- circuit boards designed for use with spring or wire contacts
- modular keyboard configuration allowed within maximum range of 96 keys: easy implementation of, e.g., 54- or 72-key units
- inputs suitable for driving from contacts other than those on a musical keyboard
- keyboard matched to controller either by software (EPROM contents) or hardware (physical connection of contacts)

work contact. The present keyboard works on the same basis.

Strike the right note with the E510

The Type E510 MIDI controller is without doubt a revolutionary integrated circuit, and has been recognized as such by many readers following the publication of the *Portable MIDI keyboard*. The plastic package with only 16 pins (Fig. 1) contains a programmed control circuit with MIDI keyboard functions normally carried out by a fast microprocessor and one or more peripheral circuits. However, the E510 also has its drawbacks and limitations: it recognizes only one split, while up to 16 can be programmed on many keyboards. Also, the E510 can send data to MIDI channels 1 and 2 only. The velocity parameter can not be geared precisely to the characteristics of the keyboard, or be given the optimum range to suit the average strike force of the user.

Contrary to the single-chip, mask-programmed E510, most microprocessor systems are 'open' which means that they may be programmed or re-programmed to include the above features. The E510, on the other hand, has the advantage of being extremely simple to use in a practical circuit. Acknowledging the fact that the vast majority of musicians working with MIDI equipment are not electronics buffs, a simple circuit is a significant factor.

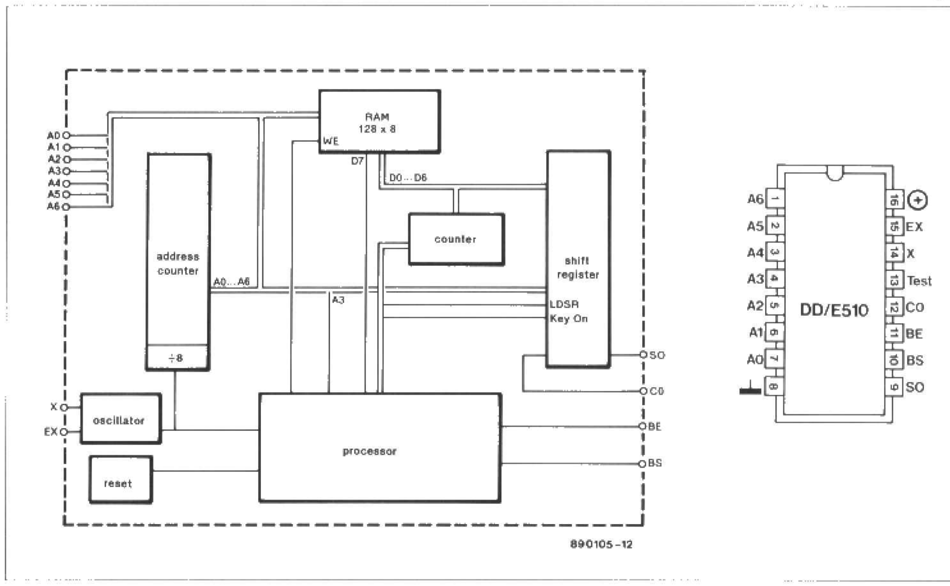


Fig. 1. Pinning and block diagram of the single-chip MIDI keyboard controller Type E510.

A number of readers have expressed their doubts and reservations about the dynamic range of the E510. These doubts are really not justified. In fact, the velocity processor in the E510 is so good that the chip is capable of distinguishing between a soft, normal and hard keystroke even when Digitast keys are used as on the *Portable MIDI keyboard* (Ref. 1). Digitast keys have tactile feedback which makes them quite unsuitable for providing velocity information, as is clearly explained in the relevant article (this is not to say that the *Portable MIDI keyboard* is touch-sensitive in the sense specified by the MIDI standard). The present MIDI keyboard is fully equipped for velocity processing, however, and the fact that it also uses the E510 is proof of our confidence in the chip.

Before studying the circuit and the contents of the transposition EPROM, get the right orientation by briefly looking at Fig. 2, the block diagram of the MIDI keyboard. Constructors of the *Portable MIDI keyboard* will easily recognize the general structure.

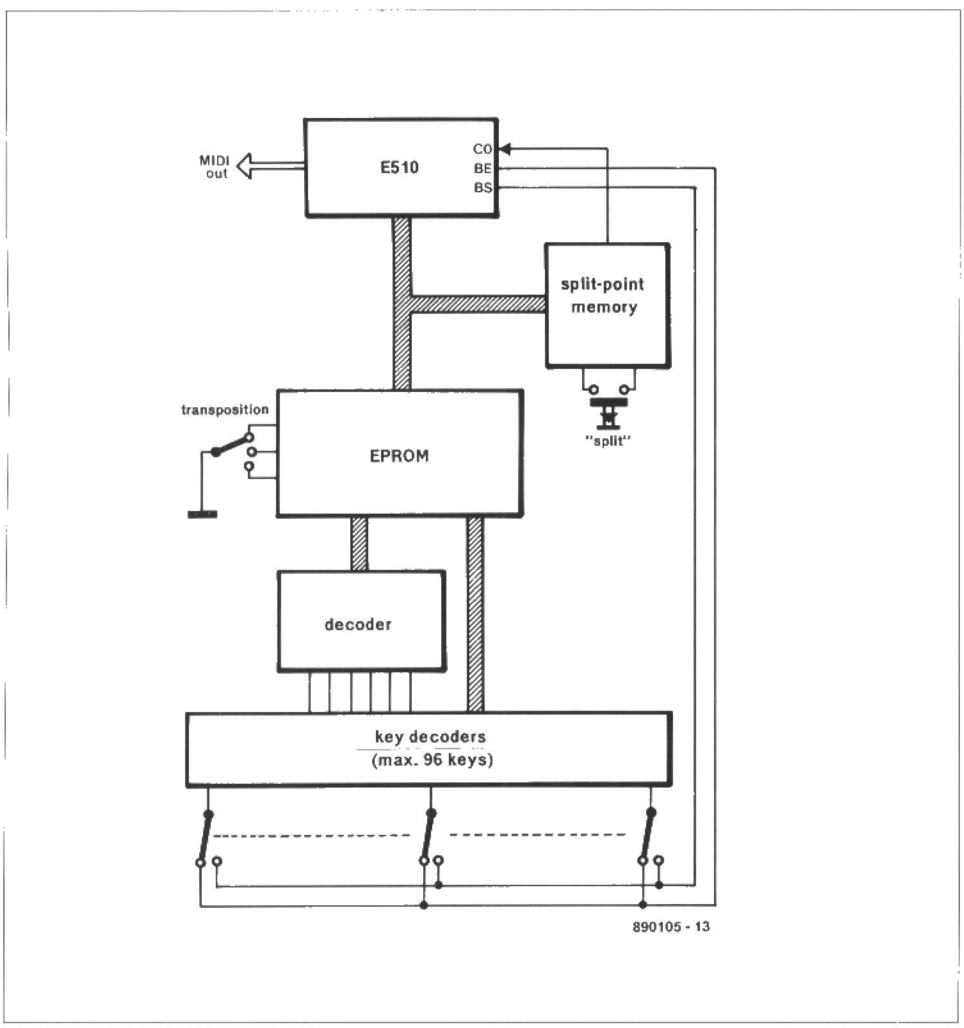


Fig. 2. Block diagram of the universal MIDI keyboard controller.

Circuit description

To avoid an unnecessary large and cluttered circuit diagram, Fig. 3 shows the (entirely theoretical) configuration of the MIDI controller with 16 keys only. The circuit diagram in fact shows only one of the possible six key decoders that may be installed. As a result of this simplification, the diagram is hardly any more complex than that of the *Portable MIDI keyboard*.

As shown by Fig. 3, each of the six key decoders is capable of addressing up to 16 key contacts, so that a maximum of 96 key contacts is available (the grand piano keyboard has 88 keys). The circuit diagram of the keyboard section in two possible versions is given in Fig. 6 (its operation will be discussed in due course).

As already stated, the basic operation of the E510 keyboard controller in the present application is similar to that in the *Portable MIDI keyboard*. Details of the key scanning mode and velocity processing are, therefore, not repeated here since these have been covered at length in Ref. 1.

The E510 has an on-board 7-bit binary counter, which provides states 0 through 127 on outputs A0 through A6. Between these outputs and the key contacts sits an

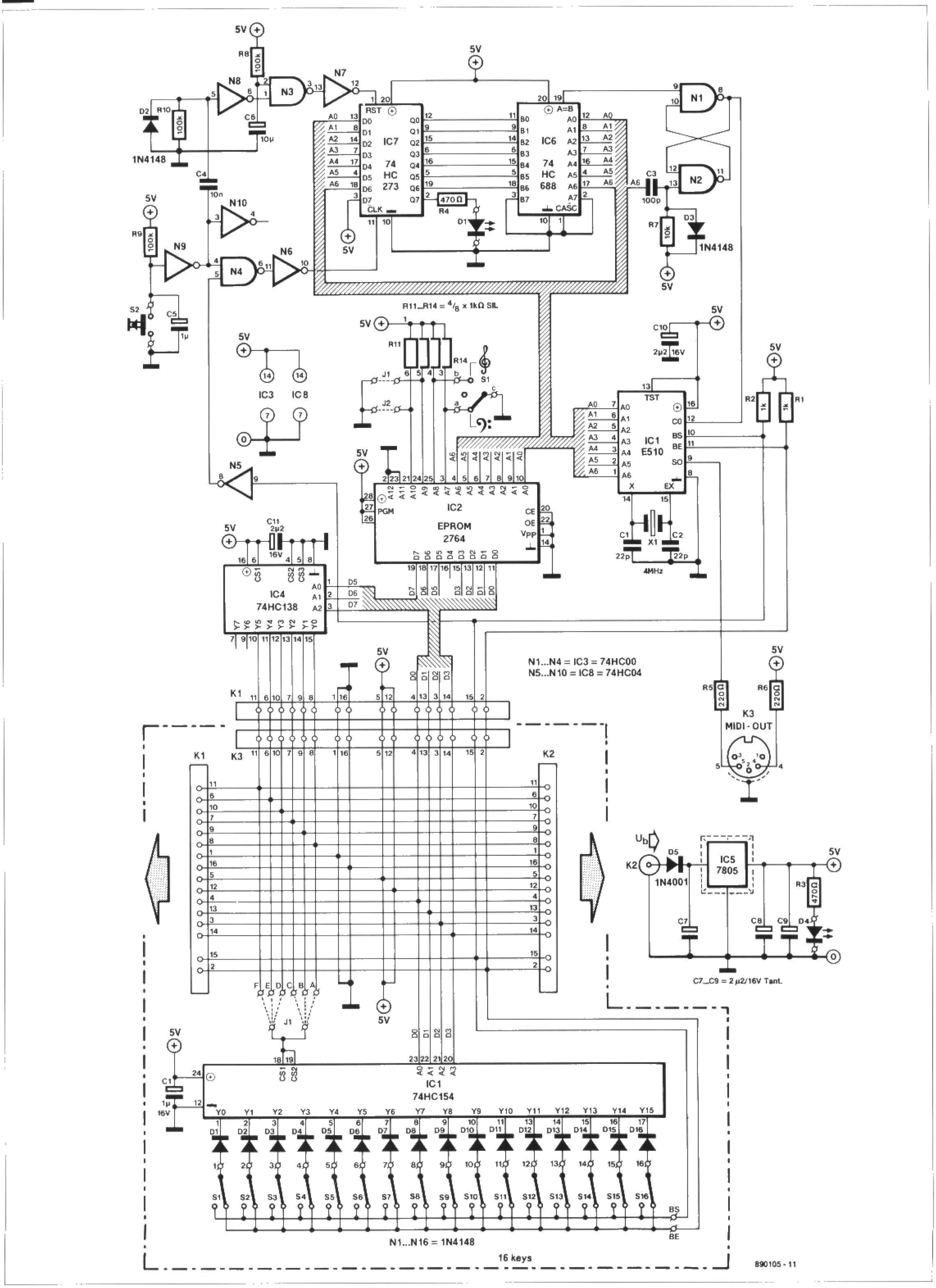


Fig. 3. Circuit diagram of the MIDI keyboard controller. For clarity's sake, only one of six 16-key decoders is shown. The configuration of the keyboard section is given in Fig. 6.

address transcoder in the form of an EPROM. This chip has two functions: first, it suppresses the E510-generated addresses corresponding to notes so low that they are inaudible, and, second, it allows the player to select up or down transposition of a section (zone) of the keyboard.

The binary values that appear at the counter outputs of the E510 are applied to the address inputs of the EPROM. The output word of the EPROM is available on 7 data lines. Of the 7 output bits, 4 carry the address of one of sixteen keys within a decoded group, and 3 the address of one of six decoders. The actual key addresses are carefully programmed values to obtain either the normal mode with no split points, or up/down transposition of the counter values supplied by the E510. The 4 least-significant data lines (LS nibble) of the EPROM are connected direct to the binary inputs of 1-of-16 decoders Type 74HCT154, which, in turn, are connected to the key contacts. The most-significant data lines of the EPROM (MS nibble) drive the address decoder, a 1-of-8 decoder Type 74HCT138, whose outputs enable the six key decoders. With the exception of the 74HCT138, the keyboard interface is basically the same as that used in the *Portable MIDI keyboard.*

The addition of the 1-of-8 decoder and some modifications to the EPROM contents make it possible to increase the number of keys to that required for a full-size MIDI keyboard. The relation between the keyboard type and the EPROM contents will be reverted to.

Split-point

Briefly, a split-point, or simply split, on a MIDI keyboard effectively splits the keyboard into two smaller keyboards, whose size in terms of keys is defined by the player. The principle is illustrated in Fig. 4. On a 6-octave keyboard, for instance, the 2 low octaves may be assigned to a bass instrument on MIDI channel 1, while the higher 4 octaves are assigned to another instrument, say, piano accompaniment, controlled via MIDI channel 2.

The top part of Fig. 3 shows the split-point programming circuit. The E510 scans the keyboard in low-to-high order, i.e., from the key producing the lowest note to the one producing the highest note. A split is simply programmed by actuating push-button S2 simultaneously with the key that defines the wanted position of the split. This action causes the address of the key to be stored in memory. The output of the split-programming circuit pulls input CO of the E510 high while the chip scans the keyboard, and a key is addressed with a number higher than that of the key that defined the split-point. The E510 responds to the high level at CO by redirecting all MIDI data to output channel 2 rather to than channel 1. When the key scanner has reached the highest key, i.e., when the E510 has passed counter state 127, the split-programming circuit is reset, and CO is made logic low again, so

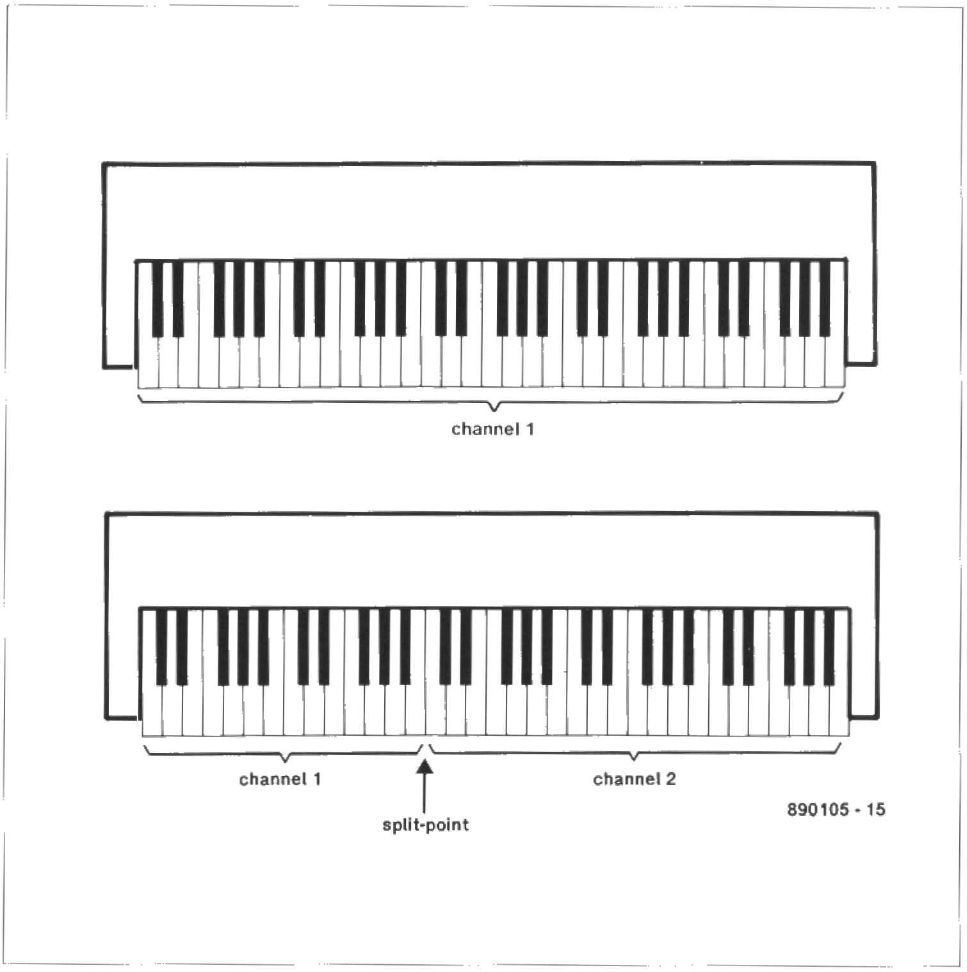


Fig. 4. Illustrating the principle of keyboard section rerouting to alternate MIDI channels, known as *split-point programming.*

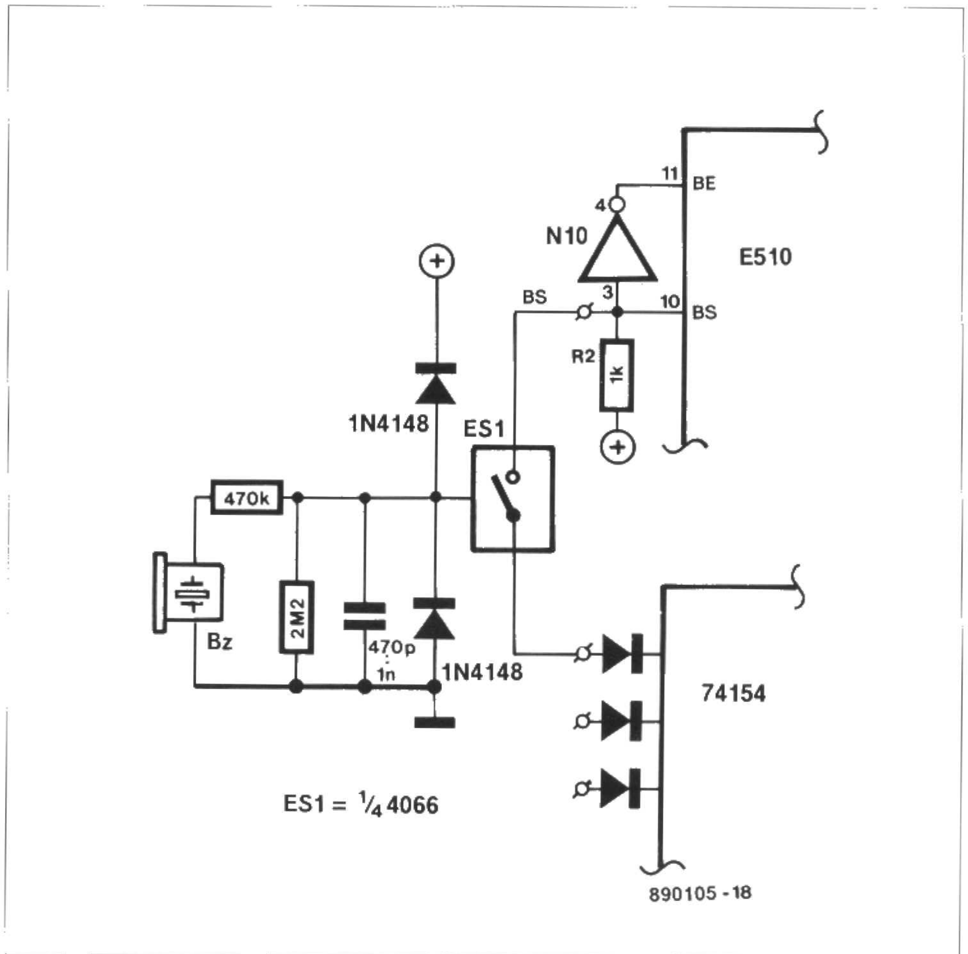
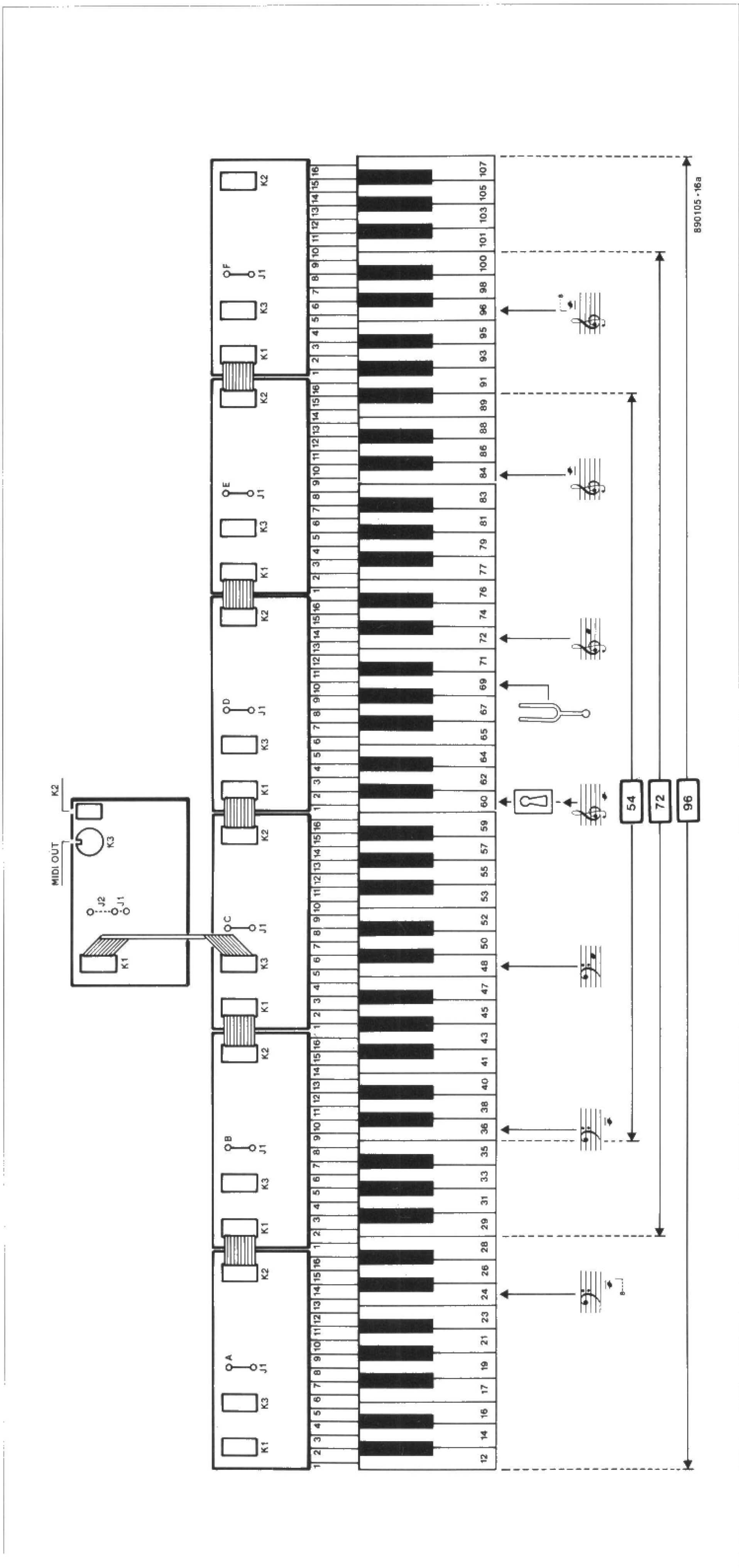


Fig. 5. Experimental percussion interface (see Ref. 2).



that MIDI data is routed to channel 1 again.

The split-programming circuit can only store a key address when line BS is low, which is the case when the pole of the addressed key reaches the work contact, and S_2 is closed. In that condition, gates N_9, N_4 and N_6 generate a positive pulse transition at the CLK input of IC_7. This octal bistable then copies the logic combination applied to its inputs, D0–D7, to its outputs, Q0–Q7. The combination forms the address of the key actuated by the player programming the split. Bit D7 does not form part of this address: it is forced logic high and causes D_1 to light, indicating that a split has been programmed.

During subsequent keyboard scan cycles, IC_6, an 8-bit comparator, compares the address stored in memory and applied to its inputs B0–B6 to that available on the address bus of the E510 and applied to its inputs A0–A6. When these addresses are equal, i.e., when the keyboard scanner reaches the key that defined the split, the bistable formed by N_1 and N_2 is set to logic 1 by output A=B of the 74HCT688 (pin 9 of N_1). Input CO of IC_1 goes logic high. At the end of the keyboard scan cycle, the bistable is reset to logic 0 by the negative pulse transition on address line A6, which drives differentiator C_3-R_7-D_3.

When input CO of the E510 is low, MIDI data is routed to channel 1. When CO is high, it is routed to MIDI channel 2. At power-on, the bistable is reset to 0 by R_7-C_3. Octal latch Type 74HCT273 is also reset at power-on with the aid of a low pulse generated by R_8-C_6 and applied to the RST input. Actuation of S_2 when no key is pressed (BS is logic 1), causes network C_4-R_{10} connected to N_8 to reset the latch also, while any previously programmed split is erased. Diode D_2 protects the input of N_2 against voltage peaks.

In practice, it is recommended to always erase an old split before programming a new one simply by pressing S_2 only.

It is possible to direct the 'low' keyboard section to the left of the split to

Fig. 6a. Configuration of an integral 96-key keyboard. Databyte 00 is loaded in the EPROM at relative address 12_{10}, or $0C_H$ counting from the start of block 180_H addresses in normal mode without transposition. A keyboard with 72 keys starting with note *F* may 'start' on the second contact of the second lowest decoder (selected with link B). Non-used contacts may be left open, or connected to the BE line to simulate the presence of rest contacts. In that case, the first decoder board, normally enabled by link A, need not be installed.
When it is desired to have, for example, 3 complete *C*-to-*C* octaves to the left of the middle C, the keyboard must start one octave lower at the *F* note corresponding to MIDI KEY 17. In that case, the board selected by link A must be installed, while the last board enabled by link F may be omitted.
A 54-key *C*-to-*B* keyboard, for instance, starts at contact S_9 of the second board.

channel 2, and the section to the right of the split to channel 1, instead of the other way around which forms the default configuration. Two possibilities exist for this modification:

- insert non-used inverter N_{10} (IC_8) in the CO line (pin 12) of the E510;
- break the connection between input CO and the output of N_1 (pin 8 of IC_3). Connect input CO to the output of N_2 (pin 11 of IC_3) instead. This modification causes an 'unsplit' keyboard to address MIDI channel 2 instead of 1 at all times.

Inverter N_{10} in IC_8 is useful when the velocity parameter is to be omitted. In that case, the rest contacts of the keys need not be connected because only the work contacts are used. Indeed, the keys need not have a rest contact at all. Line BE must, however, be forced high by the actuated BS signal, and be forced low when BS is inactive. To free the BE input, remove pull-up R_1, and connect it to the output of N_{10}, whose input is connected to BS. This modification is illustrated in Fig. 9.

Percussion enthusiasts are referred to Fig. 5, which shows an interface that allows the keyboard inputs to be driven by signals obtained from a simple beat detector built from a piezoceramic buzzer (Ref. 2).

Transposition by EPROM

The first task of the EPROM is to place the physical keyboard in the range of 128 virtual keys addressed by the E510. The controller counts from 0 to 127 irrespective of the actual number of keys connected. Without a decoder or transposition circuit, the lowest key on the keyboard would correspond to key MIDI 0. This is not vey useful because this key number belongs to a subsonic frequency. The EPROM thus allows the real keyboard to be centred around number 60 of the 128 virtual keys. This centre is formed by the middle C as illustrated in Figs. 6a and 6b..

Since enough space is left in the EPROM, the complete physical keyboard can be transposed towards the low or high end of the virtual keyboard. This is the second function of the EPROM, whose available memory capacity is, however, still not exhausted. Therefore, jumpers J_1 and J_2 are provided to give access to normally unused memory in the EPROM for the implementation of special functions.

The jumpers are normally installed so that effectively the lower quarter of the

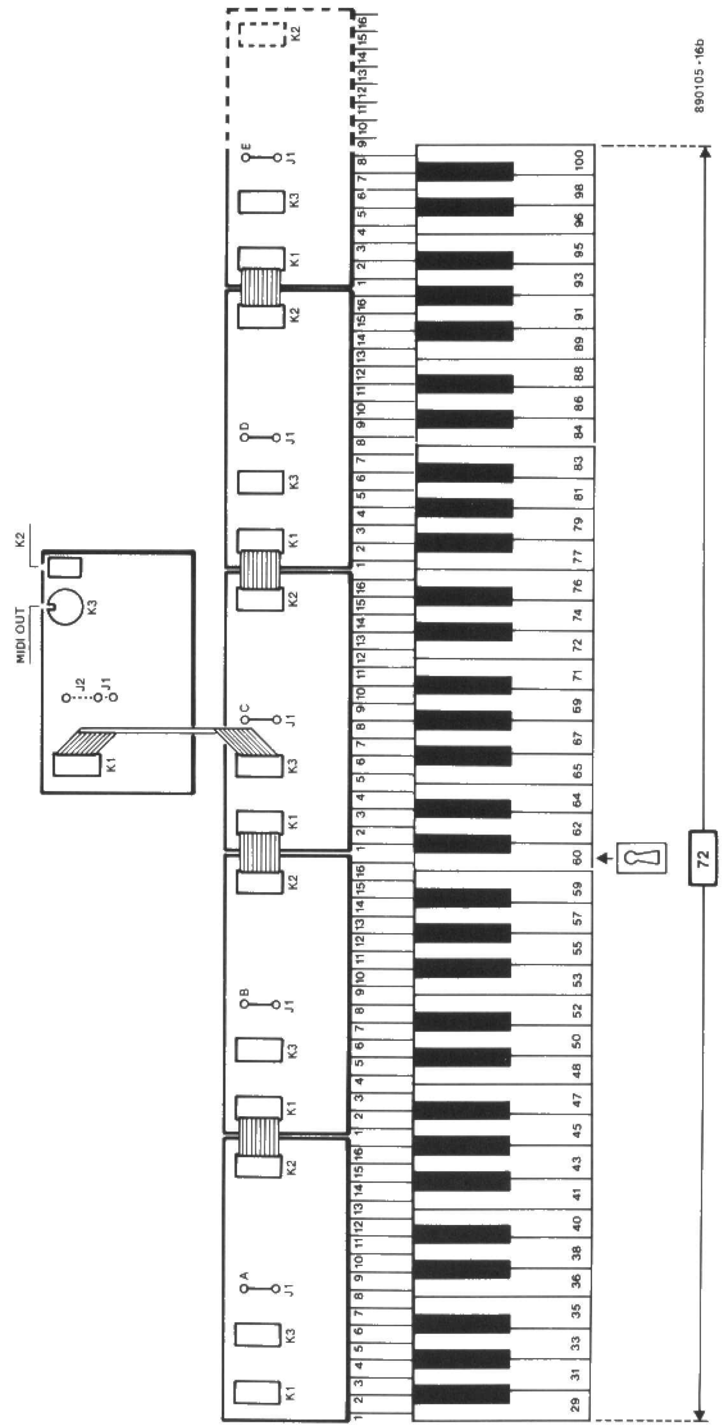


Fig. 5b. Configuration of a 72-key keyboard. The EPROM is re-programmed such that the first contact of the first decoder board corresponds to the first key of the keyboard. In normal (non-transposed) mode, databyte 00 (see Table 1) is loaded in the EPROM at relative address $1D_H$, or 29_{10}, counting from the start of block 0180_H in Table 3 (this will be given in Part 2).

Table 1.

Switch S_1 sets the logic levels on address lines A7 and A8, and so selects between normal operation, up-transposition or down-transposition:

A8	A7	A6–A0: counter 0–127
0	0	not allowed
0	1	transpose down
1	0	transpose up
1	1	normal configuration

EPROM contents for virtual keyboard with 96 notes from C to B

*C = *middle C* on the first additional line under the treble stave.

	A8	A7	A6																	
n	0	0	0	0000	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
o	0	0	0	0010	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
t	0	0	0	0020	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	0	0	0	0030	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
u	0	0	1	0040	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
s	0	0	1	0050	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
e	0	0	1	0060	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
d	0	0	1	0070	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	0	1	0	0080	00	01	02	03	04	05	06	07	08	09	0A	0B	0C	0D	0E	0F
					C		D		E	F		G		A		B	C		D	
	0	1	0	0090	20	21	22	23	24	25	26	27	28	29	2A	2B	2C	2D	2E	2F
					E	F		G		A		B	C		D		E	F		G
L	0	1	0	00A0	40	41	42	43	44	45	46	47	48	49	4A	4B	4C	4D	4E	4F
						A		B	C		D		E	F		G		A		B
O	0	1	0	00B0	60	61	62	63	64	65	66	67	68	69	6A	6B	6C	6D	6E	6F
					C		D		E	F		G		A		B	*C		D	
W	0	1	1	00C0	80	81	82	83	84	85	86	87	88	89	8A	8B	8C	8D	8E	8F
					E	F		G		A		B	C		D		E	F		G
	0	1	1	00D0	A0	A1	A2	A3	A4	A5	A6	A7	A8	A9	AA	AB	AC	AD	AE	AF
						A		B	C		D		E	F		G		A		B
	0	1	1	00E0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					C		D		E	F		G		A		B	C		D	
	0	1	1	00F0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					E	F		G		A		B	C		D		E	F		G
	1	0	0	0100	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					C		D		E	F		G		A		B	C		D	
	1	0	0	0110	FF	FF	FF	FF	FF	FF	FF	FF	00	01	02	03	04	05	06	07
					E	F		G		A		B	C		D		E	F		G
H	1	0	0	0120	08	09	0A	0B	0C	0D	0E	0F	20	21	22	23	24	25	26	27
						A		B	C		D		E	F		G		A		B
I	1	0	0	0130	28	29	2A	2B	2C	2D	2E	2F	40	41	42	43	44	45	46	47
					C		D		E	F		G		A		B	*C		D	
G	1	0	1	0140	48	49	4A	4B	4C	4D	4E	4F	60	61	62	63	64	65	66	67
					E	F		G		A		B	C		D		E	F		G
H	1	0	1	0150	68	69	6A	6B	6C	6D	6E	6F	80	81	82	83	84	85	86	87
						A		B	C		D		E	F		G		A		B
	1	0	1	0160	88	89	8A	8B	8C	8D	8E	8F	A0	A1	A2	A3	A4	A5	A6	A7
					C		D		E	F		G		A		B	C		D	
	1	0	1	0170	A8	A9	AA	AB	AC	AD	AE	AF	FF	FF	FF	FF	FF	FF	FF	FF
					E	F		G		A		B	C		D		E	F		G
	1	1	0	0180	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	00	01	02	03
					C		D		E	F		G		A		B	C		D	
N	1	1	0	0190	04	05	06	07	08	09	0A	0B	0C	0D	0E	0F	20	21	22	23
					E	F		G		A		B	C		D		E	F		G
O	1	1	0	01A0	24	24	26	27	28	29	2A	2B	2C	2D	2E	2F	40	41	42	43
						A		B	C		D		E	F		G		A		B
R	1	1	0	01B0	44	45	46	47	48	49	4A	4B	4C	4D	4E	4F	60	61	62	63
					C		D		E	F		G		A		B	*C		D	
M	1	1	1	01C0	64	65	66	67	68	69	6A	6B	6C	6D	6E	6F	80	81	82	83
					E	F		G		A		B	C		D		E	F		G
A	1	1	1	01D0	84	85	86	87	88	89	8A	8B	8C	8D	8E	8F	A0	A1	A2	A3
						A		B	C		D		E	F		G		A		B
L	1	1	1	01E0	A4	A5	A6	A7	A8	A9	AA	AB	AC	AD	AE	AF	FF	FF	FF	FF
					C		D		E	F		G		A		B	C		D	
	1	1	1	01F0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					E	F		G		A		B	C		D		E	F		G

890105 - T1

EPROM is used. Removal of one or both jumpers causes another, differently programmed, address area to be selected in the EPROM. Details on programming are given in the relevant section below.

Jumper J_1 selects one of six enable signals A–F in the key decoding circuit. There are 16 contacts to each keyboard sub-circuit. The number of decoders required depends on the number of keys available on your keyboard. Jumpers are, therefore, placed to individual requirement. Examples: a 4-octave keyboard requires at least 3 address decoders, a 64-key type 4, and a 72-key type 4½ as illustrated in Fig. 5. At least five decoders are required for 80 keys, 5½ for 88 keys, and, finally, all 6 for 96 keys.

The jumper for the first decoder (at the 'low' side of the keyboard) is marked A, the next one B, and so on, up to jumper F, which enables the decoder that reads the highest 16 keys.

The standard EPROM contents correspond to a 96-key keyboard with a tone range from C (MIDI KEY NUMBER 12) to *B* (MIDI KEY NUMBER 107). Figure 6b illustrates the fitting of a 72-key keyboard with range *F* to *E* into the 96-key range addressed by the EPROM. The actual number of keys matters very little, provided double addressing is avoided. More importantly, however, the number of the lowest key of the keyboard used must correspond to the counter value reserved for it by the E510. In other words, if, for example, a 54-key *C*-to-*F* keyboard is available, an EPROM may be used with the contents given in Table 1, but only if the lower *C* of this keyboard is connected to contact S_9 of the second decoder board, as shown in Fig. 5.

Modifying the EPROM contents to suit individual requirements is not necessary in most cases, but fairly simple on the basis of the information given below.

Programming the EPROM

The standard contents of the EPROM for a 96-key keyboard are listed in Table 1. To facilitate altering the contents, Table 2 gives the unprogrammed 'framework' which serves to document one's own EPROM contents. Table 2 can be completed by entering the actual key numbers as shown in the example of Table 3 (this will be included in next month's instalment).

Having studied the circuit diagram of the MIDI controller, it will have been noticed that output bit D4 is not used. Normally, bit 7 is not used, but here the design of the printed circuit board has forced the omission of bit 4. The upshot is that the most-significant nibble in the databyte is always nought or an even number (0, 2, 4, 6, or 8), as shown in Tables 1 and 3. Mind this simple rule when compiling and programming your own EPROM with the aid of Table 2.

Possible misgivings about the versatility of the MIDI keyboard should be dispelled by the fact that the EPROM may

hold up to 64 different keyboard configurations. Jumpers J1 and J2 allow the selection of 16 different tables. The remaining 48 are available after modifying the connections of address lines A11 and A12. Electronics enthusiasts not interested in electrophonics may like to know that the E510, in conjunction with a microprocessor, is also eminently suitable for building an advanced multi-point contact scanner.

References:

1. Portable MIDI keyboard. *Elektor Electronics* November 1988.
2. Disco drum. *Elektor Electronics* June 1984.

The construction of the MIDI keyboard will be discussed in next month's second and last instalment of this article.

Table 2.

To program the EPROM:
1. Enter '0' in the cell corresponding to the number of the lowest key on your keyboard;
2. enter the successive key numbers in ascending order, right up to the highest key.

Note: the MS nibble is either 0 or an even-numbered value. The first 128 bytes are always FF. This part of the EPROM is not accessed.

0000	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
0010	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
0020	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
0030	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
0040	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
0050	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
0060	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
0070	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF

DOWN-TRANSPOSITION (A7=1; A8=0)
*C = *middle C* on the first additional line under the treble stave.

0080	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	C		D		E	F		G		A		B	C		D	
0090	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	E	F		G		A		B	C		D		E	F		G
00A0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
		A		B	C		D		E	F		G		A		B
00B0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	C		D		E	F		G		A		B	*C		D	
00C0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	E	F		G		A		B	C		D		E	F		G
00D0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
		A		B	C		D		E	F		G		A		B
00E0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	C		D		E	F		G		A		B	C		D	
00F0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	E	F		G		A		B	C		D		E	F		G

UP-TRANSPOSITION (A7=0; A8=1)
*C = *middle C* on the first additional line under the treble stave.

0100	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	C		D		E	F		G		A		B	C		D	
0110	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	E	F		G		A		B	C		D		E	F		G
0120	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
		A		B	C		D		E	F		G		A		B
0130	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	C		D		E	F		G		A		B	*C		D	
0140	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	E	F		G		A		B	C		D		E	F		G
0150	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
		A		B	C		D		E	F		G		A		B
0160	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	C		D		E	F		G		A		B	C		D	
0170	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	E	F		G		A		B	C		D		E	F		G

NORMAL (A7=A8=1)
*C = *middle C* on the first additional line under the treble stave.

0180	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	C		D		E	F		G		A		B	C		D	
0190	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	E	F		G		A		B	C		D		E	F		G
01A0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
		A		B	C		D		E	F		G		A		B
01B0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	C		D		E	F		G		A		B	*C		D	
01C0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	E	F		G		A		B	C		D		E	F		G
01D0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
		A		B	C		D		E	F		G		A		B
01E0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	C		D		E	F		G		A		B	C		D	
01F0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	E	F		G		A		B	C		D		E	F		G

EVENTS

IEE Meetings

2 June — Application of microprocessors to the control/monitoring of transmission and distribution equipment.
5 June — The status of electrical machines in transport and industry.
6 June — Integrated optics.
7 June — Passive intermodulation products in antennas and related structures.
9 June — Electronic filters.
19–30 June — Photographic and Engineering Models Exhibition.
20 June — Outside broadcast and satellites.
21 June — Power station practice for developing countries.

Readers are also advised that each year the IEE organizes a range of conferences, vacation schools, technical seminars and workshops on subjects within the fields of electrical, electronic and control engineering, and computing.

Further information on these, and many other, events from IEE • Savoy Place • LONDON WC2R 0BL • Telephone 01-24- 1871.

APRS the **International Exhibition of Professional Recording Equipment** will take place on **7–9 June** at Olympia 2, London W14. Further information from Association of Professional Recording Studios • 163A • High Street • RICKMANSWORTH WD3 1AY • Telephone (0923) 772907.

EUUS, the **European Unix User Show** will be held on **6–8 June** at the Alexandra Palace Pavilion, LONDON N22. Further information from EMAP International Exhibitions Ltd • 12 Bedford Row • LONDON WC1R 4DU • Tel. 01-404 4844.

PPE, the **Power Plant Exhibition** will take place on **20–23 June** at the National Exhibition Centre, Birmingham. Further information from Swan House Special Events Ltd • Holly Road • HAMPTON HILL TW12 1QQ • Tel. 01-783 0055.

The **Networks '89** Exhibition and Conference will be held at the NEC, Birmingham on **6–8 June**.

The **Software Tools '89** Exhibition and Conference will be held on **13–15 June** at the Wembley Conference & Exhibition Centre, London.

The **IT in Government** Exhibition will take place at the Wembley Conference & Exhibition Centre, London on **13–15 June**. The associated conference will take place on the same dates at the Hilton Hotel, Wembley.

The **European Satellite Broadcasting** conference will be held at the QEII Centre, London, on **22–23 June**.

Details of these events from Blenheim Online Ltd • Blenheim House • Ash Hill Drive • PINNER HA5 2AE • Telephone 01-868 4466.

SMT/ASIC, an international Exhibition and Conference on Surface Mount Technology and Application Specific ICs will be held at Nurnberg on **20–22 June**. Details from Mesago • Telephone +49 711 618075.

Computer North will take place at Manchester on **6–8 June.** Details from Cahners Exhibitions Limited • Chatsworth House • 59 London Road • TWICKENHAM TW1 3SZ • Telephone 01-891 5051.

A **Computer, Software and Electronics Fair** will be held at Cologne on **15–18 June**. Further information from the German Chamber of Commerce, London. Telephone 01-930 7251.

SEMI's **Third European Forecast Conference**, which has as its theme **"Facing the Challenges of the 90s"** will be held at the Hotel Martinez, Cannes, on **19–21 June.** Details from SEMI European Secretariat • CCL House • 59 Fleet Street • LONDON EC4Y 1JU • Telephone 01-353 8807.

Scrambler Chip Keeps Phonecalls Private

Protection against eavesdropping on open-transmission telephone conversations when confidentiality is required is provided by a speech scrambler, developed by Marconi Electronic Devices, which offers a low-cost method of providing effective security and good voice quality.

The "DVS Speech Encryption Processor" is based on a CMOS scrambler IC designed for use in narrow-band analogue speech encryption. With the addition of RAM and minimal control circuitry, the IC uses the Time Division Multiplexing (TDM) technique to encrypt and decrypt two analogue speech channels for duplex or simplex operations.

TDM involves dividing, in the time domain, the digitally converted speech waveform into small sections known as frames. These frames of digital samples are reversed in time and then converted back to an analogue signal. This encrypted signal is then transmitted across the radio link. When this signal is received, the process is repeated and clear speech is once again obtained.

Low-cost Video Production Aids

Three compact and user-friendly accessories for the low-cost production of professional-quality video have been introduced by Unitron Products to enable the small production facility to produce creative and imaginative video programmes.

The "CXL01 Timebase Corrector" reduces the expense of signal stabilization essential in VCR mixing.

The "AVD2.5" is a high-quality two-channel, ten-output stereo audio/visual distribution amplifier that is equally well suited to network-distribution or tape-duplication applications, as well as Y-C tape formats.

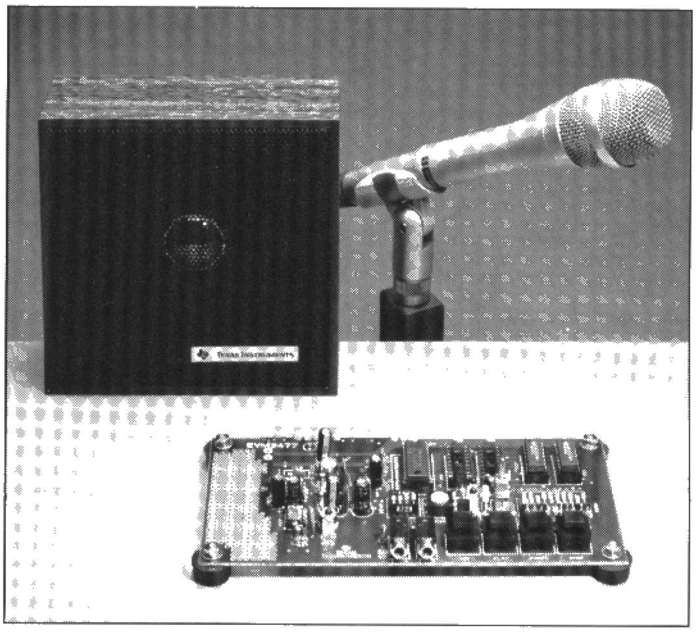
The "VPA 03 Dubmaster" is a versatile colour corrector and processor suitable for operation with standard VHS, the latest S-VHS and the professional Umatic formats.

Details from Unitron Products • 183 Sunningvale Avenue • Biggin Hill, Kent TN16 3TL • Telephone (0959) 71313 • Telex 95358.

APPLICATION NOTES

The contents of this column are based on information obtained from manufacturers in the electronics industry, or their representatives, and do not imply practical experience by *Elektor Electronics* or its consultants

Voice recorder from Texas Instruments



For many years now, the most popular means of analogue recording and playing back of audio signals has been the cassette recorder. But even here, digital techniques are beginning to make inroads. True, available material allows only relatively short recording times, but for a number of applications, for instance, telephone answering machines, advertising messages, memory aids, alarm installations, and so on, it is perfectly usable.

A new IC from Texas Instruments, the TMS 3477, is intended as basis for such equipment. Apart from RAM, all necessary functions are available on the chip. The block diagram of a possible system is shown in Fig. 1. The IC may be operated in two different ways. The simpler is by means of a four-position keyboard, of which the keys assume the functions corresponding to those normally available on a cassette recorder. The other method is via a computer. Dynamic RAMS instead of cassette tapes are used as recording medium. If you want to listen to something different, you insert a different bank of DRAMS or make a new recording.

A modified form of continuously variable slope delta modulation (CVSD) is used in the TMS 3477 for the quantization (digitization) of the audio signals. This type of modulation used with DRAMS has the important advantage of requiring only simple connections between the TMS3477 and the DRAMS.

The principle of CVSD is shown in Fig. 2. The analogue signal, u_x, is compared with u_y, a signal that increases or diminishes only slowly. Whether u_y increases or diminishes depends on u_x, which in its turn depends on the difference between u_x and u_y. The digital signal u_x thus contains information on the analogue signal. Since it is a digital signal, it may be stored in a memory.

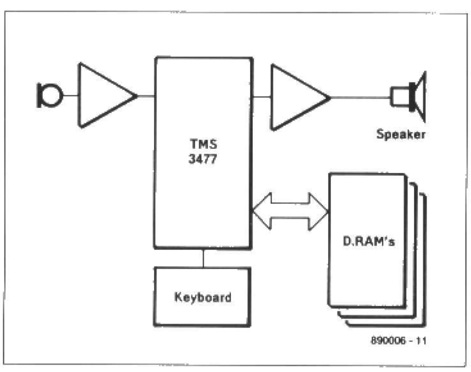


Fig. 1. A recorder system based on the TMS3477 and its block diagram.

Another advantage of delta modulation is that the integrator of the modulator may be used also as demodulator. Signal u_y then serves as the output signal.

The integrator (which is indispensable for delta modulation) is built up in the TMS 3477 rather differently from what you might expect. It is constructed from an adder and a digital-to-analogue converter. The adder is the real integrator, since, in this case, integrating is nothing more than increasing the preceding result by 1 (if u_x is high) or reducing it by 1 (if u_x is low). The converter has been added to translate the digital content of the adder into an analogue signal, u_y, which is either fed to the comparator or, during playback, to the output.

Several of these stages may be recognized immediately in Fig. 3. First, there are the comparator, the data latch, the adder and the digital-to-analogue converter that form the delta modulator. To these are added two further integrators to enable the speed with which u_y can change is matched to the signal level. This greatly improves the quality of the sound.

The remainder of the chip consists of the necessary control logic for the external memories and the host interface via which the TMS 3477 is controlled.

An experimental circuit diagram for a complete recorder system is shown in Fig. 6. The TMS 3477 contains a mode register that defines the execution mode. This register is programmed at the power-on reset via the address outputs of the DRAMS (AP0–AP9, where AP stands for Address/Program), which serve as temporary input during the reset procedure.

Since the AP pins serve as inputs and outputs, the logic levels for initializing the IC MUST be applied via pull-down resistors (R_1–R_{10}) – pull-up resistors have already been provided on board the chip. Table 1 summarizes the functions that may be realized via these pins.

The type of RAM that will serve as memory for recording is set via pins AP0 and AP1. There is a choice of 3: TMS 4164 (64 Kbit); TMS 4256 (256 Kbit); and TMS 4C1024 (1 Mbit). Up to two RAMS (only of the same type) may be connected. Whether one or two are used is indicated via AP2.

Switches S1 and S2 further extend the

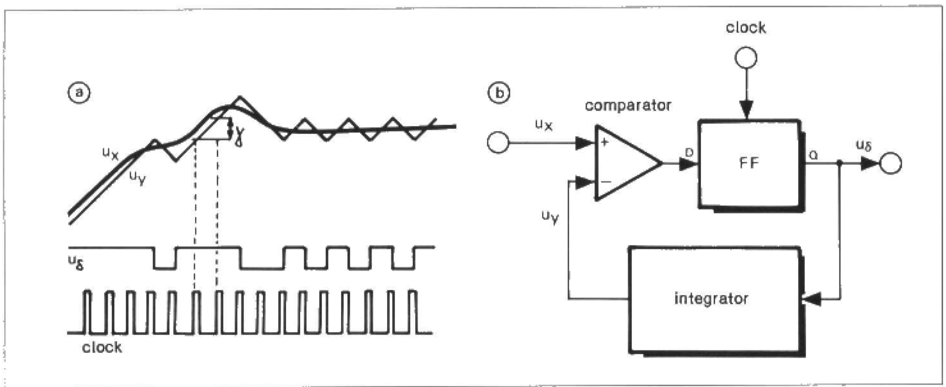


Fig. 2. The principle of continuously variable slope delta modulation (CVSD)

possibilities of the RAMs. When S2 is open, it is possible to select either of the two RAMs by S1. This enables two different phrases to be selected—PH(rase)1 and PH(rase)2. With S2 closed and S1 in position PH1 (obligatory), it is possible to record and playback **one** phrase which may, however, be twice as long as either PH1 or PH2.

The next setting refers to the length of playback period. This may be given a fixed value equal to the maximum, of which more later. With variable playback period, (too long) intervals at the end of a recording may be prevented. If after a

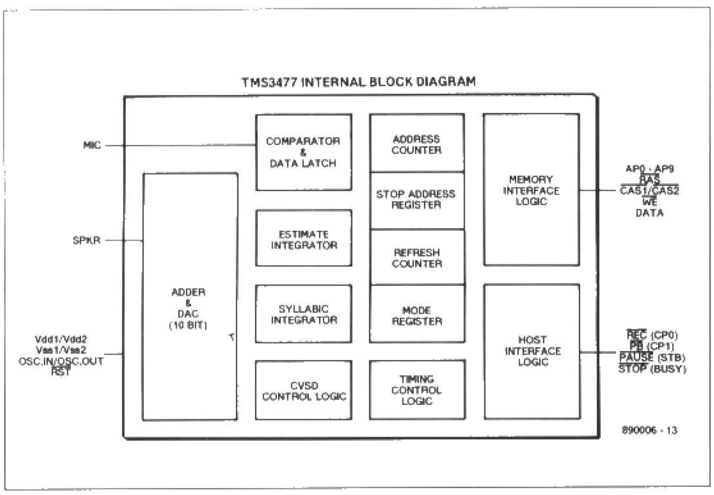
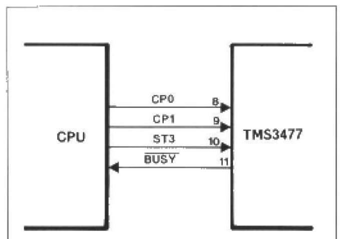


Fig. 3. Internal structure of the TMS 3477.



CP1	CP0	STB	command
0	0	⎍	pause
0	1	⎍	play
1	0	⎍	record
1	1	⎍	stop

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Fig. 4. One way of connecting the TMS 3477 to a computer.

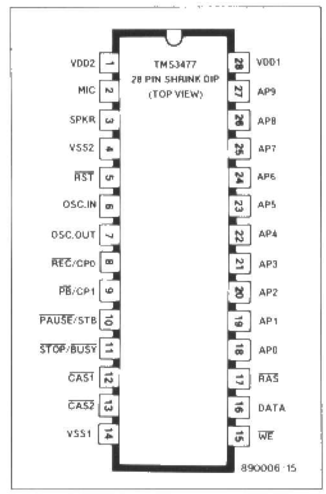


Fig. 5. Pin-out of the TMS 3477.

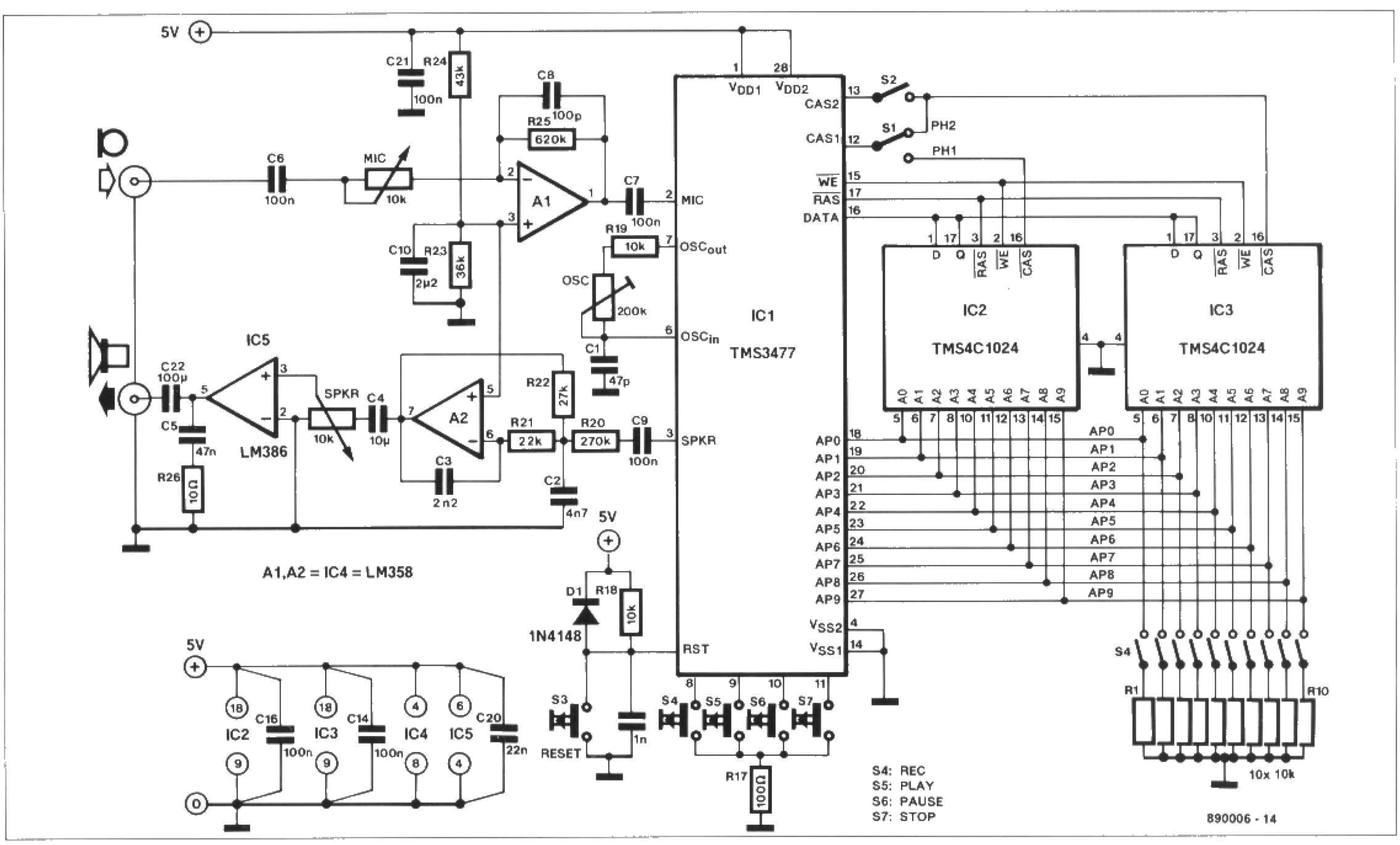


Fig. 6. Circuit diagram of an experimental voice recorder based on the TMS 3477.

function	address/program lines									
	0	1	2	3	4	5	6	7	8	9
TMS4256 DRAM	1	1	x	x	x	x	x	x	x	x
TMS4164 DRAM	0	1	x	x	x	x	x	x	x	x
TMS4C1024 DRAM	1	0	x	x	x	x	x	x	x	x
1 DRAM installed	x	x	1	x	x	x	x	x	x	x
2 DRAMs installed	x	x	0	x	x	x	x	x	x	x
variable playing time	x	x	x	1	1	x	x	x	x	x
fixed playing time	x	x	x	0	1	x	x	x	x	x
inhibit cyclic recording	x	x	x	x	1	x	x	x	x	x
cyclic recording	x	x	x	x	0	x	x	x	x	x
keyboard-interface	x	x	x	x	x	1	x	x	x	x
CPU-interface	x	x	x	x	x	0	x	x	x	x
sample frequency 32 kHz	x	x	x	x	x	x	1	1	x	x
sample frequency 16 kHz	x	x	x	x	x	x	0	1	x	x
sample frequency 64 kHz	x	x	x	x	x	x	1	0	x	x
inhibit data compression	x	x	x	x	x	x	x	x	1	x
data compression	x	x	x	x	x	x	x	x	0	x
inhibit recording monitor	x	x	x	x	x	x	x	x	x	1
exhibit recording monitor	x	x	x	x	x	x	x	x	x	0

1 = no pull-down resistor
0 = with pull-down resistor
x = don't care

Table. 1. The TMS 3477 contains a mode register that defines the execution mode. This register is programmed at the power-on reset via input pins AP0–AP9. These pins are also used as outputs to address the external DRAMS. The type of external DRAMS used is programmed via these pins like the mode of interfacing the chip with either a keyboard or a microprocessor. This table is used for memory and interface selection and defining the type of use of the chip.

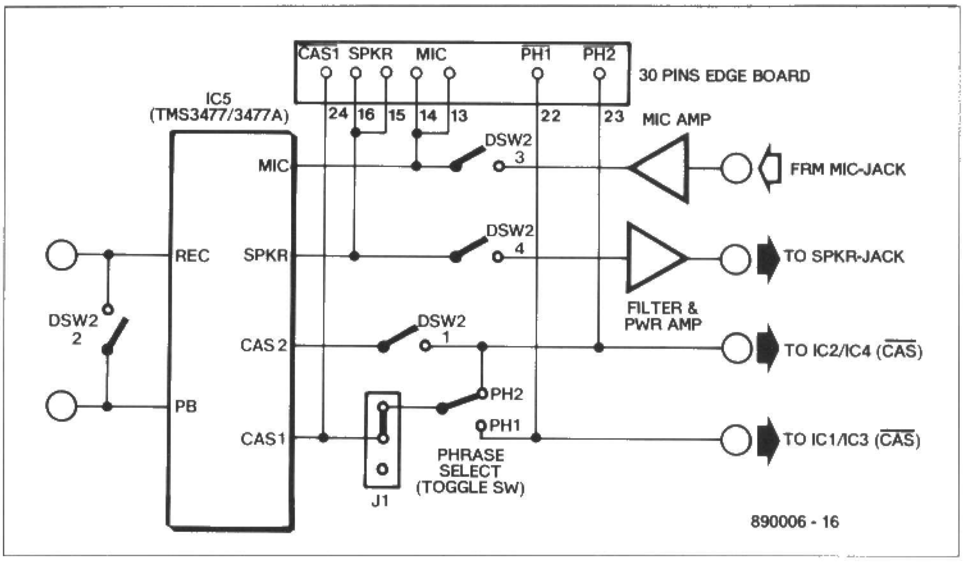


Fig. 7. Line change switches

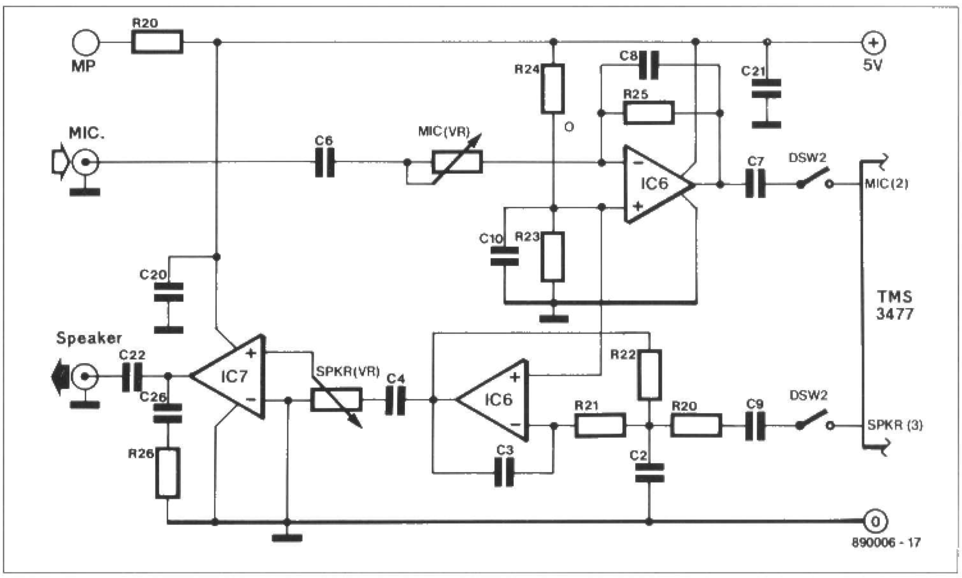


Fig. 8. Microphone and loudspeaker analogue interface.

recording the stop key is pressed, the memory address in which the last sample is stored is retained and this serves as stop address during playback later.

Another method is cyclic recording, which is set by AP4. With this method, the TMS 3477 continues recording until the stop key is pressed. Since with that method the memory will be full after a certain time, the new data is written over the old. The beginning and the end of the recording are thus 'floating around' the memory as it were. The memory therefore always contains the last section of the recorded audio signal, which is useful in, say, a dictating machine.

The type of interface via which the TMS 3477 is controlled is selected by AP5. If the keyboard is selected, the voice recorder becomes a manually controlled stand-alone unit. In this application, four switches are connected to the four interface inputs. The function of these speaks for itself.

Controlling the TMS 3477 via the CPU interface offers a number of possibilities, since the CPU allows the realization of a variety of ancillary functions, such as data transmission between two voice recorders or the storing of data in a large memory with the possibility of calling up several messages on command.

Control is effected via those pins of the IC that are also used for the keyboard interface. The functions of those pins are totally different then, however. There are two Command Port lines (CPO and CP1), a data strobe (STB) and a busy signal (see Fig. 4).

A high level on the strobe line indicates that a new command must be executed. Which demand is indicated by CP0 and CP1.

The busy signal enables the processor to check whether the TMS 3477 is in operation to prevent any unncessary breaks in recording or playback.

The sampling frequency is set via pins AP6 and AP7. Depending on the desired quality of the sound one of three available frequencies may be selected.

The duration of playback may be calculated from the sampling frequency and the memory capacity and is

memory capacity / sampling frequency

From this relationship it follows that the minimum playback time is 1 second (64 Kb; 64 kHz) and the maximum playback time is 131 s (2 Mb; 16 kHz).

A facility afforded by the digital integrator is data compression. This, in spite of its name, is a form of expansion of the audio signal. In this mode, bits are multiplied by 4 (that is, shifted to the left by two bits) before they are applied to the digital-to-analogue converter. In this way, soft recordings are reproduced much louder, albeit with a resolution of only 8 bits. This mode can not be used when recording, therefore, because this would cause a severe deterioration of the sound quality.

The last function, recording monitor, is set via pin AP9. It enables listening in during the recording.

Finally, it should be noted that the TMS 3477 is not housed in the usual DIL package, but in one with a much smaller grid (0.070" = 1.78 mm)

Source: The "TMS 3477 solid-state voice recorder" by Philippe Clement • Texas Instruments.

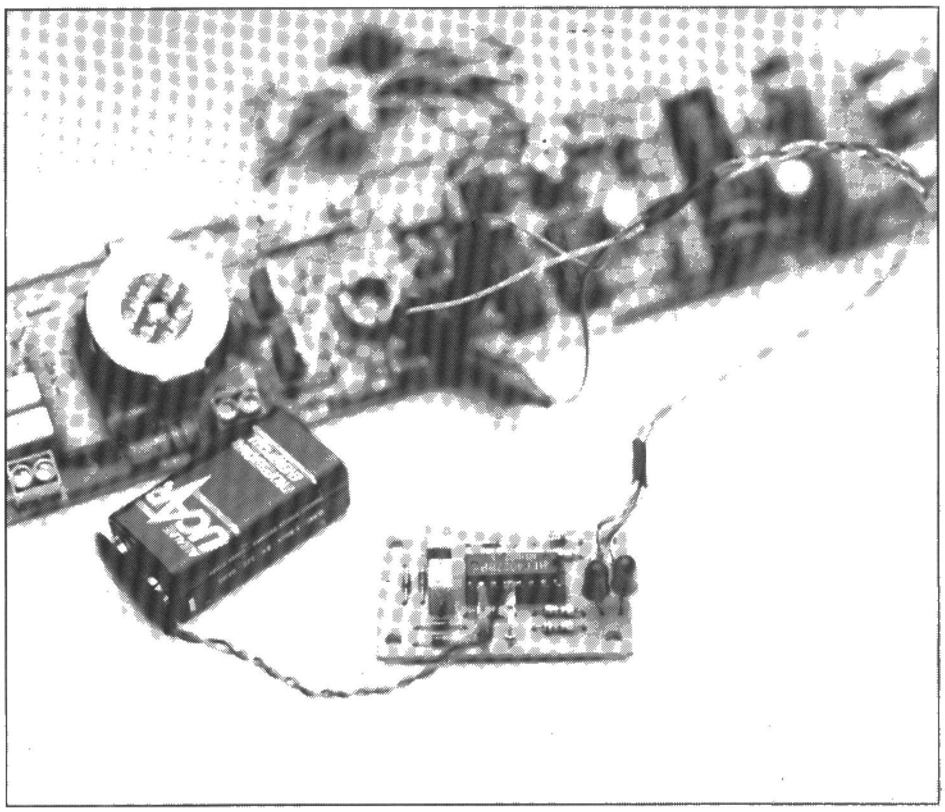
IN-CIRCUIT TRANSISTOR TESTER

by A. Rigby

In electronic troubleshooting a transistor is generally not above suspicion until it responds correctly to the usual diode-tests with an ohmmeter. Before these simple test can be performed, however, the transistor must be removed from the circuit. Experience teaches us that this operation is time-consuming as well as possibly harmful to the PCB and the rest of the circuit in a good many cases, while it offers no guarantee that the cause of the malfunction will be found.

The super-simple and inexpensive good/faulty indicator described here tests almost any transistor in circuit. A further useful feature of the tester is its built-in npn/pnp indication.

The circuit shown in Fig. 1 is straightforward and based on low-cost components. The central part is a dual J-K master/slave bistable Type 4027, IC_1, of which one section, IC_{1a}, is configured as a multivibrator. The frequency of the symmetrical output signal is set to about 100 Hz by R_1-R_2-C_1-C_2. This signal is applied direct to the input of the second bistable, IC_{1b}, which supplies the transistor under test (TUT) with two complementary-phase signals, Q and $\overline{Q}$, which have a frequency of 50 Hz.
In the absence of a TUT, current limiter R_5 passes a current through one of the LEDs, D_8 or D_9. These are connected in anti-parallel and light alternately because of the complementary drive signals supplied by the bistable. Because the LEDs are turned on and off at a rate of 50 Hz, they appear to light virtually constantly to the human eye.



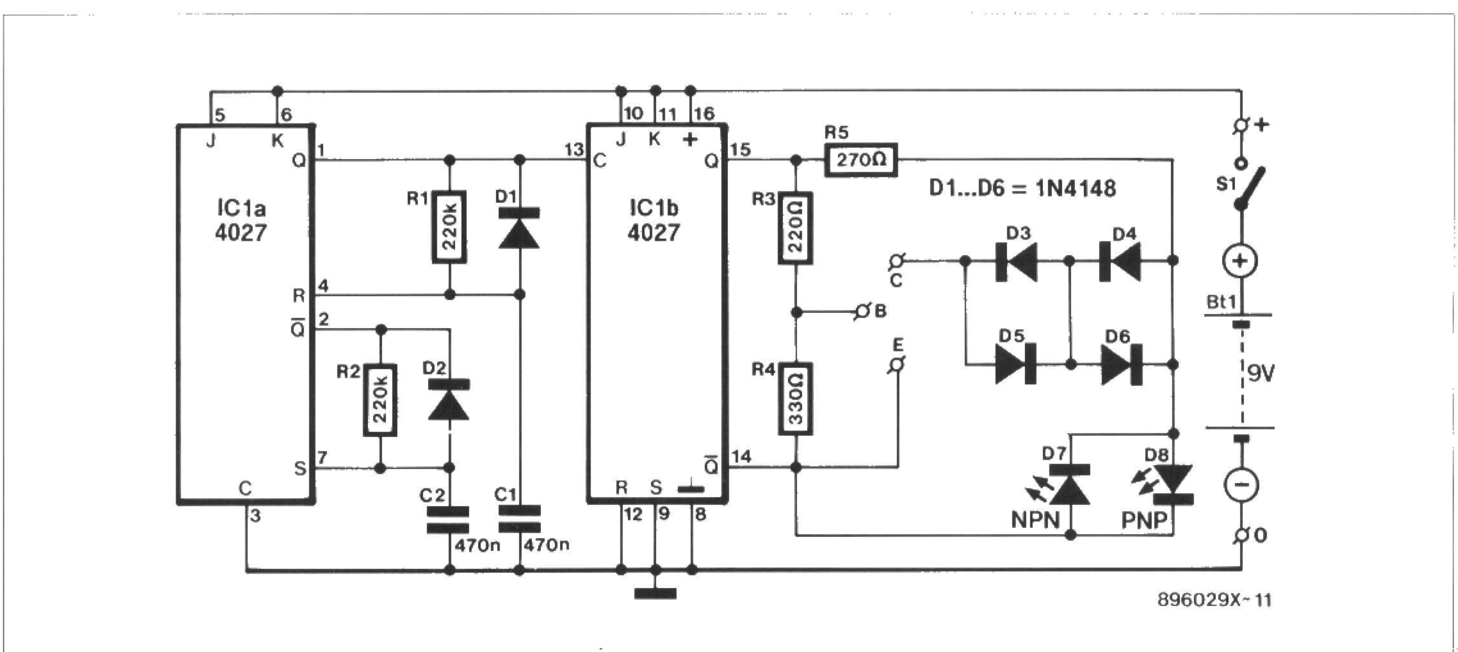


Fig. 1. Circuit diagram of the simple in-circuit transistor tester.

Bistable outputs Q and $\overline{Q}$ are connected to a potential divider, R_3-R_4. The voltage at junction R_3-R_4, $U_H/2$, is applied to the base of the TUT.

A correctly functioning npn TUT connected to test terminals B, C and E is switched on via D_3 and D_4 when Q is high and $\overline{Q}$ low, since the base is positive with respect to the emitter. Both LEDs then remain off: D_8 because it is effectively short-circuited (the drop across an intact collector-emitter junction is about 0.1 V), and D_7 because it is reverse-biased in that condition. When the bistable toggles, however, the transistor is turned off, so that D_8 is reverse-biased, and D_7 lights. The situation is reversed if a correctly functioning pnp TUT is connected: D_8 then lights while D_7 remains off.

Spotting defective transistors

Defective transistors typically have either a short-circuited or a broken collector-emitter junction. In the first case neither diode lights because of the continuous short across them. A broken c-e junction gives the same visual indication as the absence of a TUT: the LEDs light alternately.

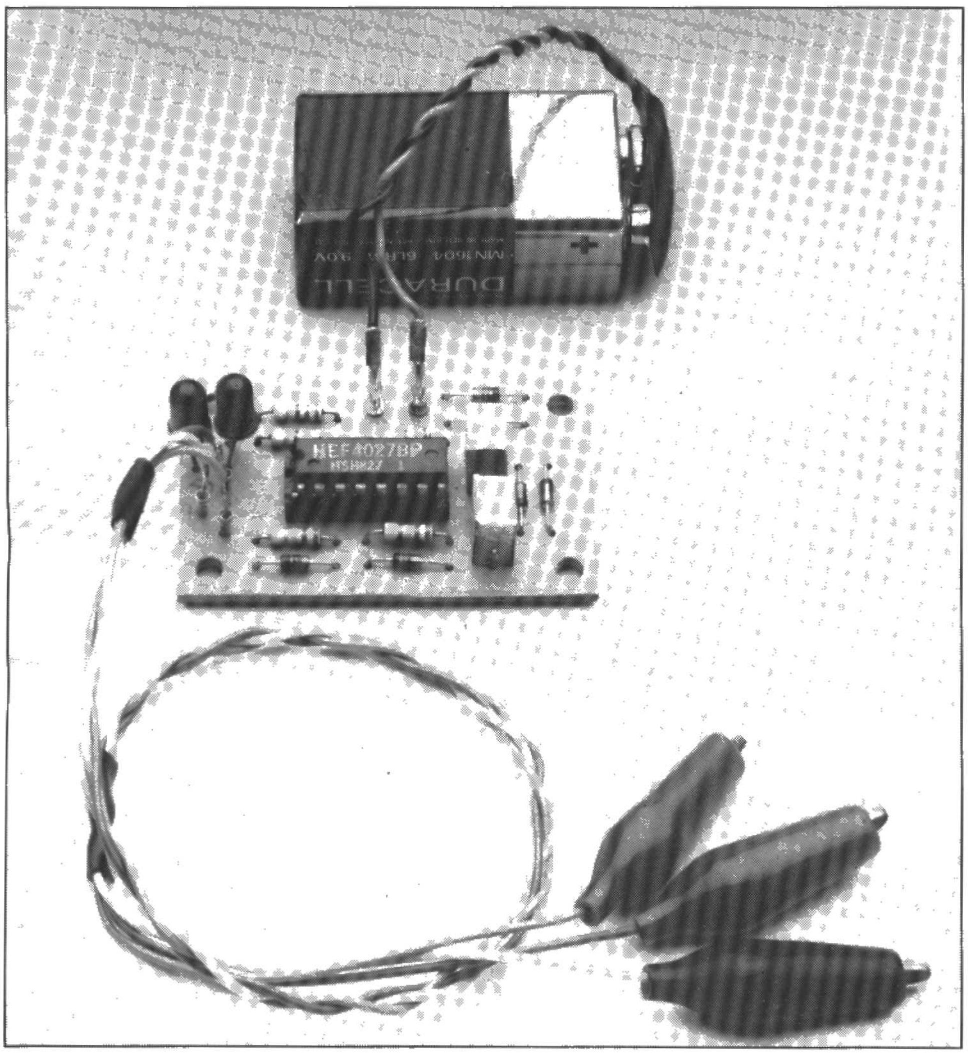
Diodes D_1-D_4 are included to prevent the tester giving an 'OK' indication with a transistor that has a base-to-collector or base-to-emitter short. This leaves only one semiconductor junction in the transistor, which then acts as a diode.

Depending on the logic state of the bistable, either D_1-D_2 or D_3-D_4 drop about 1.2 V, which is added to the drop across the collector-emitter junction of the TUT. A correctly functioning and conducting TUT has a typical c-e drop of about 0.1 V. Added to the 1.2 V introduced by the conducting pair of diodes, this voltage is not high enough to cause the turning on of the (red) LED that should remain off when the transistor is switched on. Therefore, only one LED lights: the indication is 'OK'. This changes, however, if the TUT has either of the above short-circuited junctions, since then the c-e drop becomes 0.6 V rather than 0.1 V. The resulting total drop of about 1.8 V (1.2+0.6 V) across the LEDs causes these to light simultaneously: the indication is 'faulty'.

Summarizing the above, transistors that are good are marked by only one LED (pnp or npn) lighting. All other indications (both LEDs on or off simultaneously) point to a faulty device.

Construction

The small printed-circuit board designed for the transistor tester is populated per the Parts List and the overlay shown in Fig. 2. The completed board is then installed in a plastic case with battery compartment. The tester is connected to the TUT with three flying wires with miniature, coloured and sleeved, crocodile clips. The 'on' push-button, npn/pnp indicator LEDs and, optionally, a transistor test socket, are mounted on to the front panel.



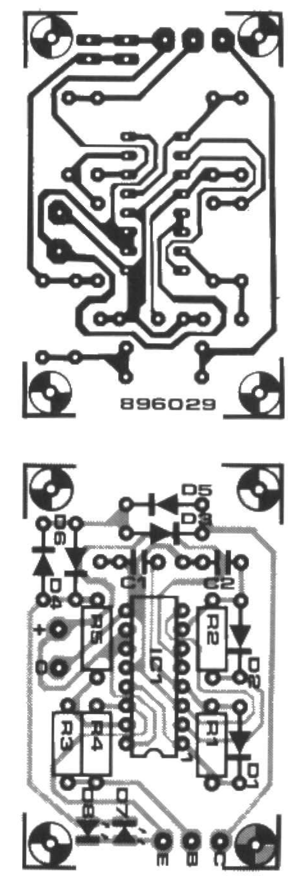


Fig. 2. True-size track layout and component mounting plan of the printed-circuit board for the transistor tester.

Parts list

Resistors (±5%):
R_1;R_2=220K
R_3=220R
R_4=330R
R_5=270R

Capacitors:
C_1;C_2=470n

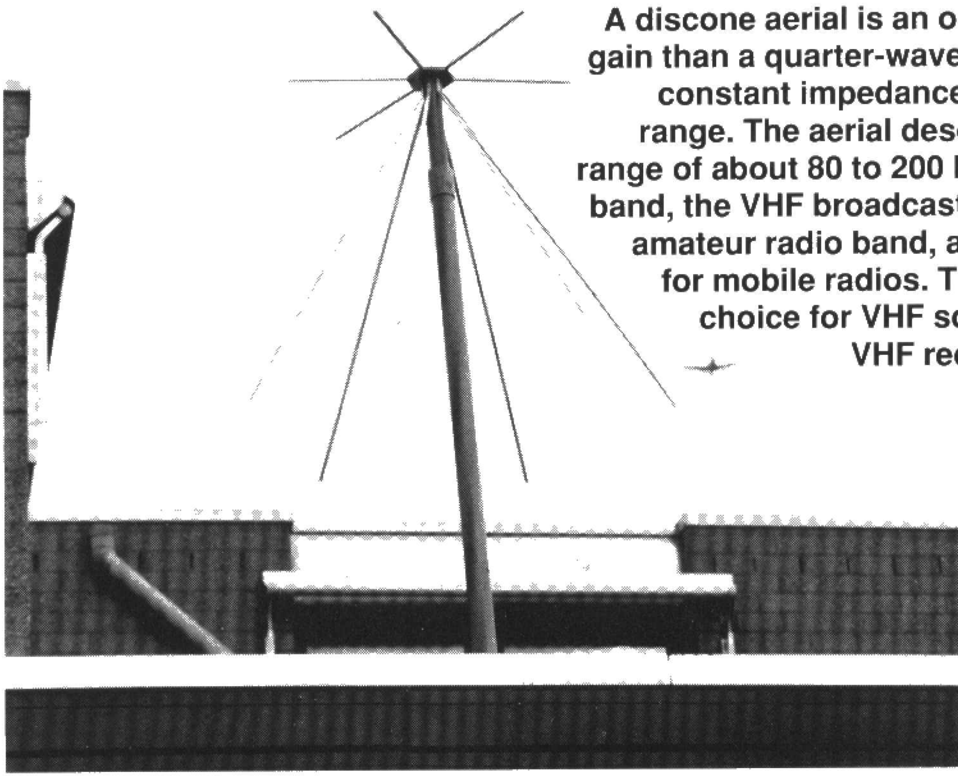
Semiconductors:
D_1...D_6 incl.=1N4148
D_7;D_8= red LED; dia. 3 mm
IC_1=4027

Miscellaneous:
S_1= push-to-make button SPST.
Bt_1= 9 V PP3 battery.
PCB Type 896029 (not available through the Readers Services).

WIDEBAND DISCONE AERIAL FOR VHF RECEIVERS

by J. Bareford

A discone aerial is an omni-directional radiator with less gain than a quarter-wave Marconi aerial, but with virtually constant impedance over a relatively wide frequency range. The aerial described here covers the frequency range of about 80 to 200 MHz, which includes the VHF-low band, the VHF broadcast band, the air-band, the 2-metres amateur radio band, and the frequency range reserved for mobile radios. The aerial is, therefore, the perfect choice for VHF scanning receivers and the AM/FM VHF receiver published earlier this year.



Most scanning receivers are capable of receiving signals over a relatively wide frequency range: 80 to 470 MHz with some gaps is not uncommon. Unfortunately, it is practically impossible to use a single aerial for the full frequency range covered by such a receiver. Most omni-directional aerials for use in the VHF band have a bandwidth of about 10 MHz only. The gain of the discone aerial at any frequency within its pass-band is lower than that of the dipole reference dimensioned for that frequency. The disadvantage of the lower gain is, however, made good by the much greater bandwidth, which is a must for scanning receivers.

Marconi aerial and ground-plane

The discone aerial is derived from the so-called Marconi-aerial, which is essentially a vertical rod or wire with a length corresponding to a quarter of the wavelength of the radiated signal. The Marconi aerial does not work unless it is placed above a relatively large conductive surface (ground plane). For VHF frequencies, the aerial will have to be installed as high as possible, requiring the ground plane to be 'taken up' also. This is achieved by simulating it with the aid of three or four grounded wires as shown in Fig. 1. The quarter-wave rod is also commonly used on cars, where the roof functions as the ground plane.

The impedance of the standard Marconi aerial with the rod and ground plane wires at right angles is 36 Ω. Fortunately, this uncommon value is simple to increase to about 50 Ω by increasing the angle between the vertical radiator and the ground plane wires (see Fig. 1b).

Disc-shaped radiator

The vertical radiator of the Marconi aerial forms a parallel tuned circuit composed of an inductor (the vertical rod) and a capacitor (between the rod and ground plane). Like any other tuned circuit, the aerial has a resonance frequency, f_{res}.

$$f_{res} = 1 / 2\pi \sqrt{LC}$$

The actual values of the inductive and capacitive component are fairly difficult to calculate. It will be clear, however, that the self-inductance increases with the length of the radiator, and the capacitance with its effective area.

In practice, a relatively large equivalent capacitance results in a larger bandwidth of the aerial. This means that the bandwidth may be increased by using a thicker radiator. To ensure that the aerial resonates at the required frequency, however, the self-inductance must be lowered at the same time. This is simple to achieve

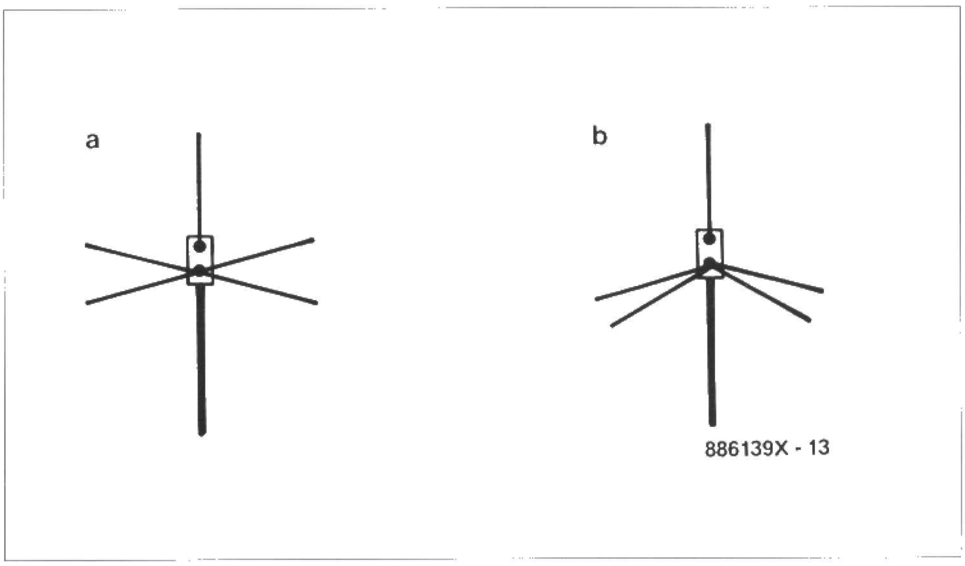


Fig. 1. The ground-plane aerial is a quarter-wave Marconi type with 3 or 4 metal rods below the radiator. These rods form a ground plane, and enable the aerial to be mounted in an elevated position. The impedance is determined by the angle between the radiator and the ground rods.

by reducing the length of the radiator.

Summarizing the above, the bandwidth may be increased by increasing the diameter of the radiator, which then requires a corresponding length reduction. The discone aerial is based on this principle, but the practical operation may not be evident at first because horizontally mounted bars or rods are used instead of a single, vertical, radiator. Like the sloping rods of the ground-plane aerial, the horizontal rods at the top of the discone aerial form a conductive plane for the frequencies involved. In effect, the diameter of the radiator is increased to the extent that the length is reduced to a few millimeters, the thickness of the rods.

Mechanical work

The discone aerial is fairly simple to make if some elementary mechanical skills are mastered. All parts are screwed together, so that welding or brazing is not required.

The construction drawing of Fig. 2 shows the overall structure, with reference numbers for all individual parts. These numbers may be found back in Table 1 and the Parts List.

The six ground elements are made from 1 m long, 10 mm wide and 5 mm thick aluminium rods. Drill two holes at one end of each bar as shown in Fig. 3. These holes are partly drilled to 7.5 mm to accept the heads of countersunk screws. Draw a line at 35 mm from the top of each rod, and use a vice and a protractor to bend it at an angle of 30°.

The six radiator elements (numbers 2 in Fig. 2) are simpler to make from 4 mm thick aluminium rods as shown in Fig. 3. The six elements are clamped between two hexagonal plates, whose dimensions are shown in Figure 4. Scale these drawings to real size to make templates for cutting and drilling the plates. The top plate (number 3) has twelve 3 mm holes for countersunk screws. Twelve 4 mm holes are drilled in the lower plate to accept threaded bushes. Where these are unobtainable, drill 3 mm holes, and use normal M3 bolts and nuts.

The small fibre plates are cut to discs of a diameter that corresponds to the inside diameter of the PVC tube. Both discs have through holes as well as holes in the side, as shown in Fig. 5. Drill 2 mm holes in the sides and tap them to provide M3 threads.

The last part is the PVC tube. The overall length is uncritical, and may exceed the indicated 250 mm (Fig. 6). A total of 19 holes is drilled in the wall of the tube — 7 holes are countersunk. Where the 12 self-locking bolts are unobtainable, drill 3 mm holes for the ground rods to enable the use of standard 3 mm bolts and nuts.

Ready for assembly

The prepared parts are held available for assembly in the following order (refer to Fig. 2):

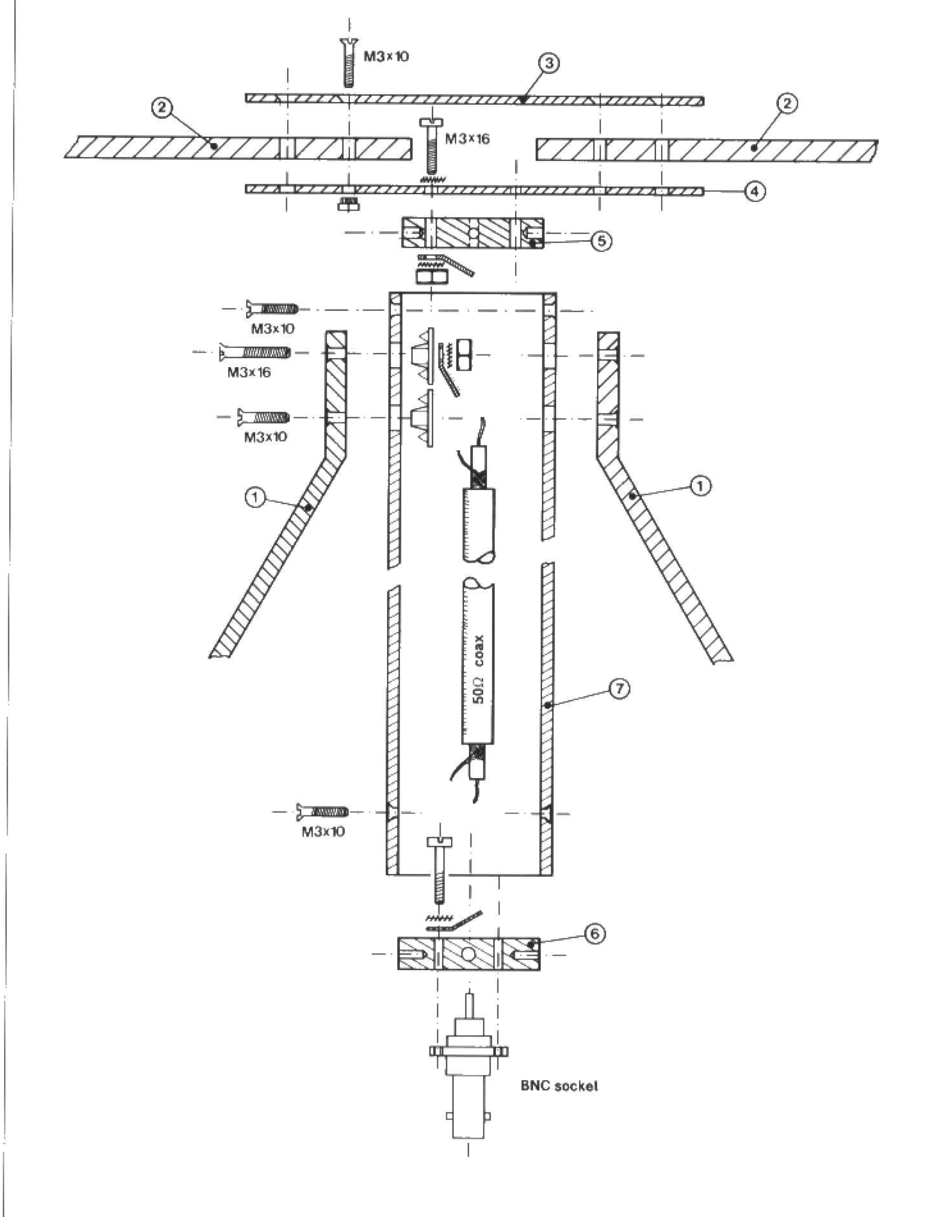


Fig. 2. General construction drawing of the discone aerial.

Table 1.

Part designation	Part description
1	6 ground elements 900×10×5 mm with 30° bend
2	6 radiator elements 300×10×4 mm
3	hexagonal top plate with 12 holes for countersunk screws
4	hexagonal bottom plate with 12 holes for M3 thread bushes, and 3 holes for M3 screws
5	man-made fibre or perspex disc for radiator
6	man-made fibre or perspex disc for BNC socket
7	PVC drainage tube

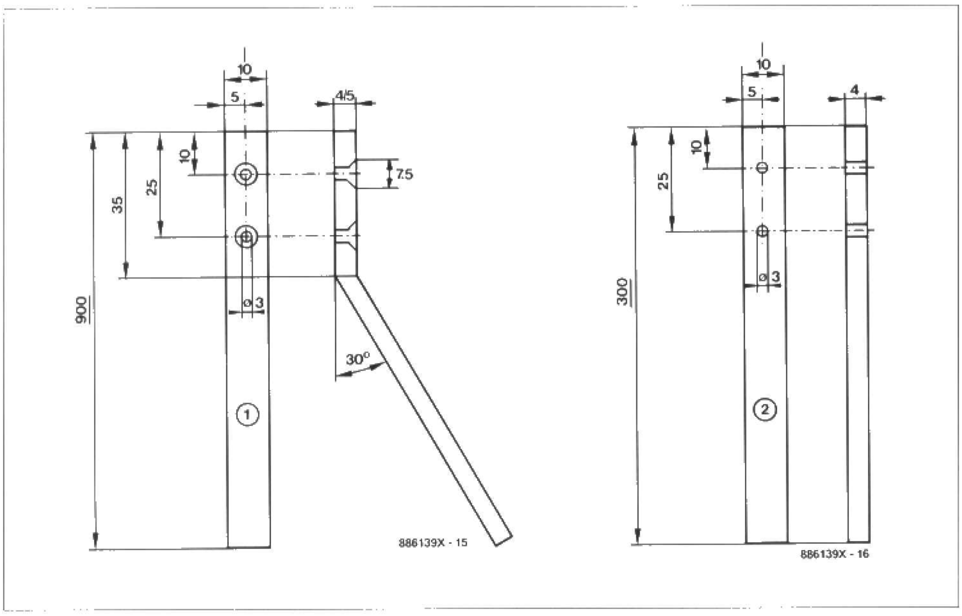
1. Use a vice to press the threaded bushes into the lower plate (4).

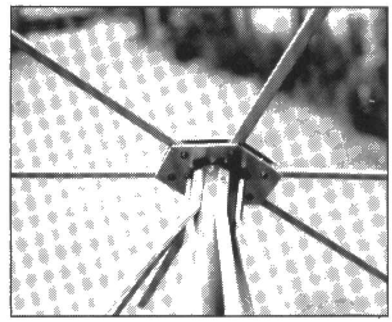
2. Screw the lower plate (4) on to the fibre plate (5) with the aid of three cylinder head screws. Use a spring washer at both ends of the screw, and insert a solder tag under one of the washers.

3. Use the countersunk screws to mount the six radiator elements (2) and the top plate (3) on to the lower plate (4).

4. Secure the six ground rods (7) on to the PVC tube (1). Use self-locking bolts, six M3×10 screws for the lower hole and six M3×16 screws for the upper hole.

5. Mount a solder tag on to each of the six screws at the upper side, and connect the tags with solid wire. Use an ohmmeter or continuity tester to check whether the





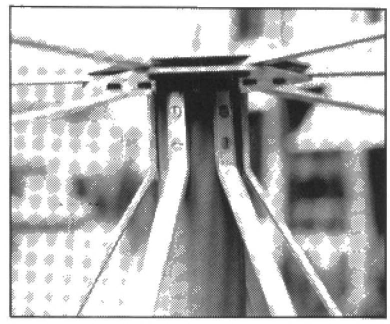


Fig. 3. Cutting and drilling details of a ground rod (left) and a radiator element (right).

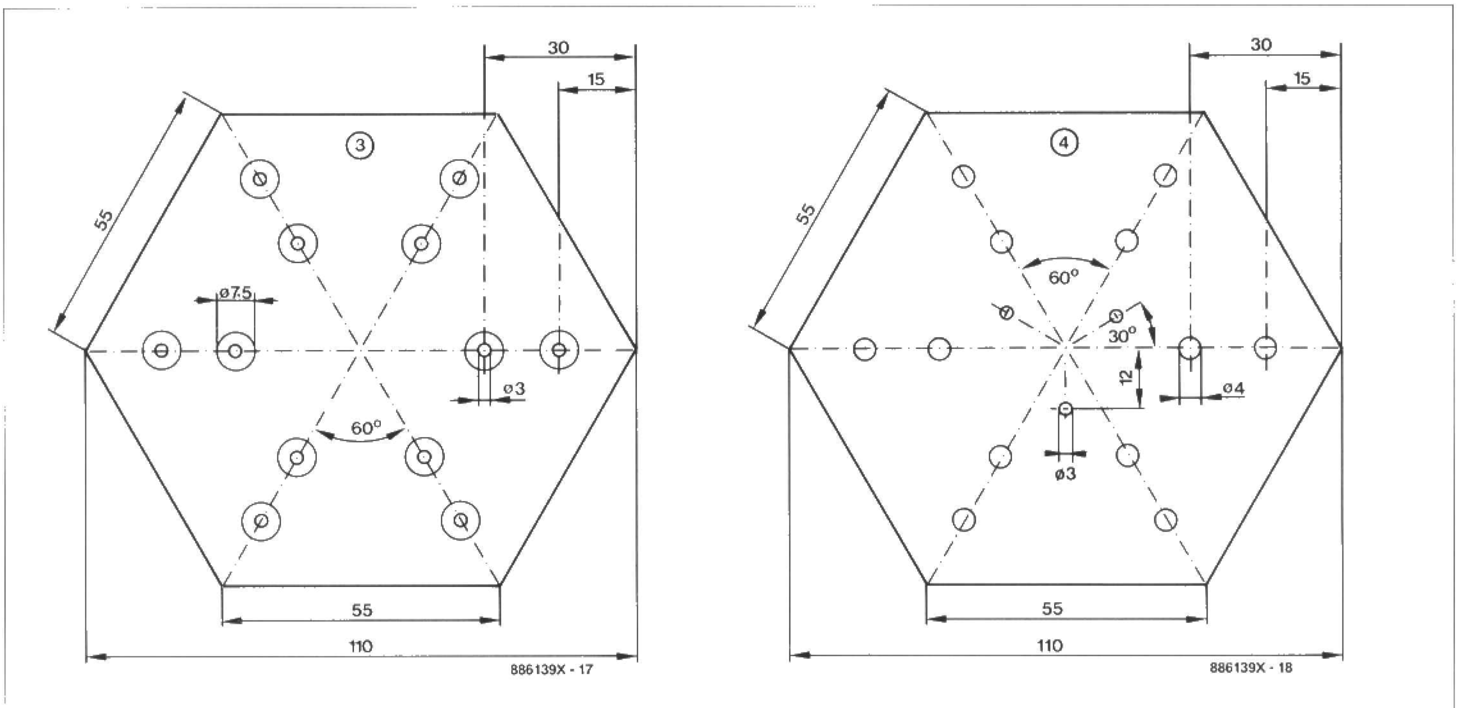


Fig. 4. All holes in the top plate (left) must be drilled to receive screws with countersunk heads. The construction of the lower plate is shown in the right-hand drawing. If threaded bushes are not available, drill 3 mm holes instead to fit M3 bolts and nuts.

ground rods are in electrical contact.

6. A piece of coax cable which is 10 cm longer than the PVC tube is stripped at both sides over a distance of about 3 cm. Insert the cable into the PVC tube and solder the shielding braid to the tags on the ground rods. Connect the core to the radiator. Secure the fibre disc (5) in the PVC tube.

7. Use the ohmmeter or continuity tester to check that there is no short-circuit between the braid and the core of the coax cable.

8. Mount the BNC socket in part (6). Many BNC sockets have threaded holes in the flange. Unfortunately, this threading is not to metric standards, so that a set of four special screws is invaluable. If such screws are not available, the threading can

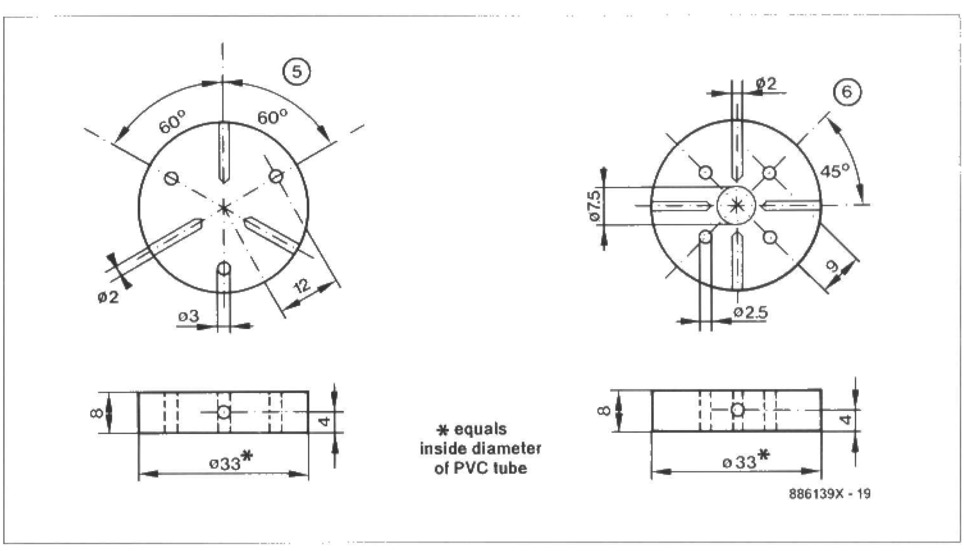


Fig. 5. The two fibre or perspex discs are cut and drilled as shown here for securing the radiator (5) and the BNC socket (6).

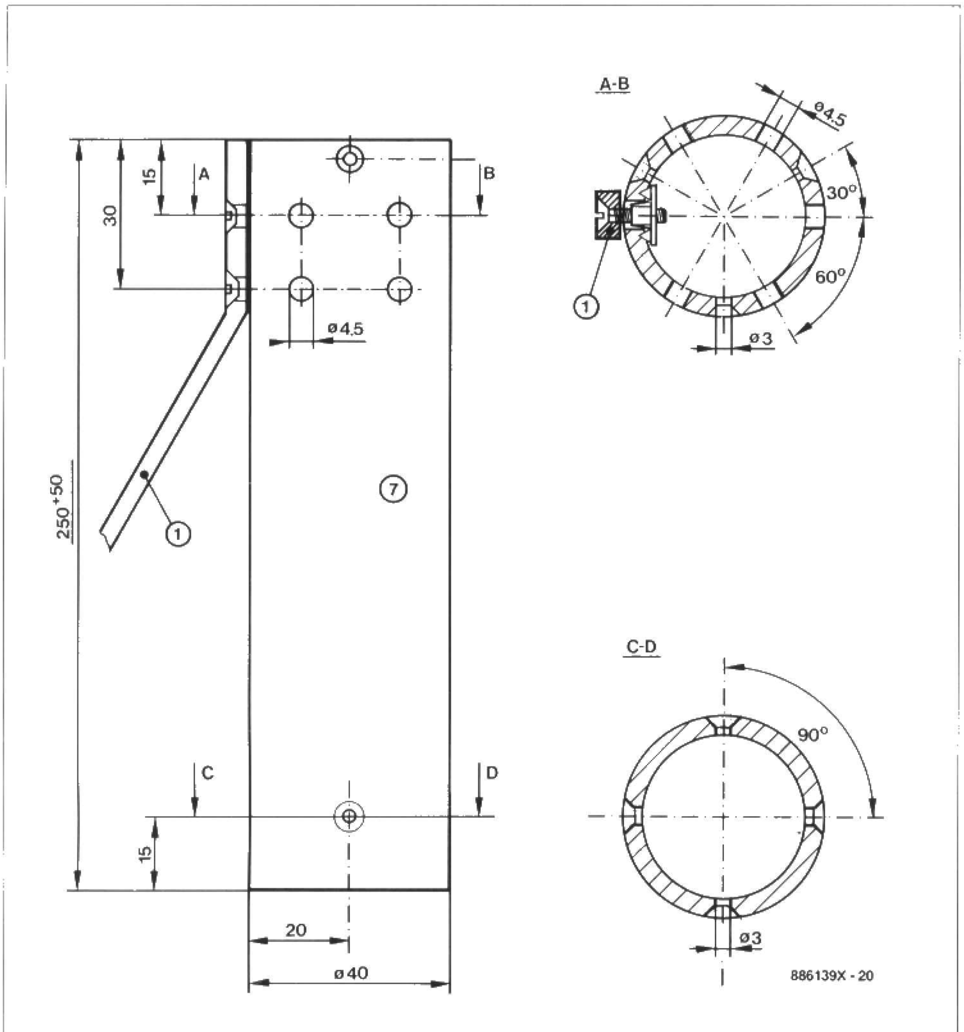


Fig. 6. Drilling and mounting details of the PVC tube, which holds all elements of the aerial together. Mark the holes with a scriber to centre the drill.

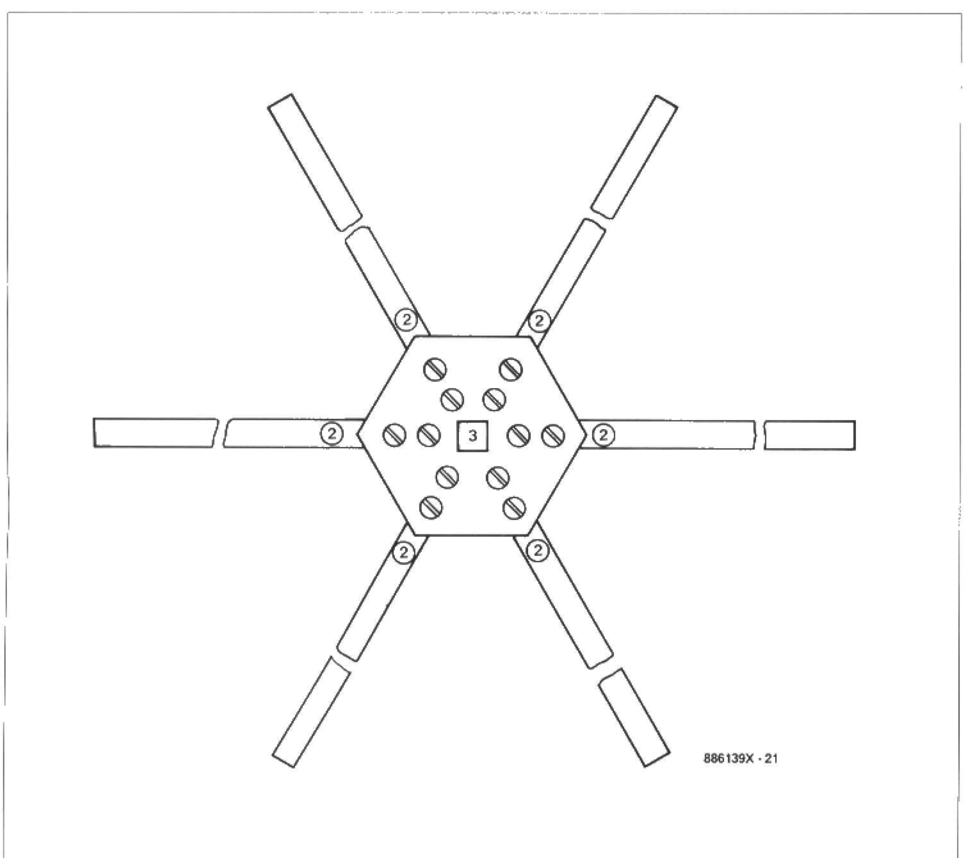
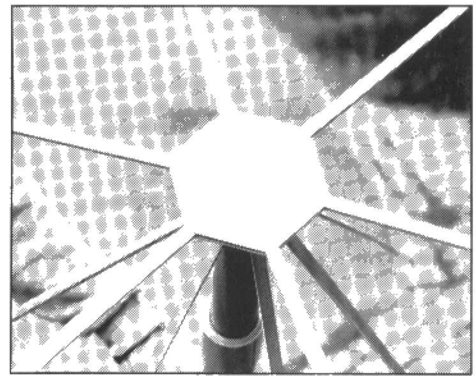


Fig. 7. Top view of the radiator element.

be drilled out so that M3 screws and nuts may be used. Do not forget to fit a solder tag under one of the screw heads.

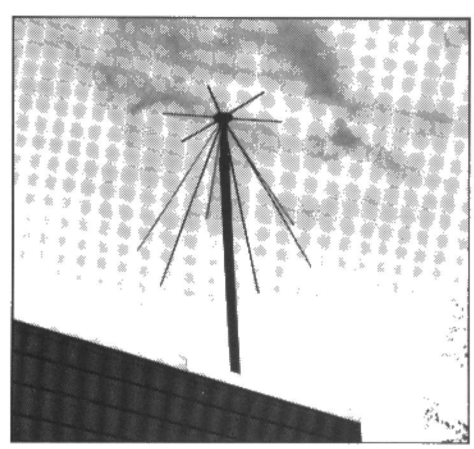
9. Solder the coax cable to the BNC socket, and secure the fibre disc (6) in the PVC tube. Once again check for a short-circuit, this time direct on the aerial connection.



The discone aerial is ready for installing on to the roof. Be sure to install it clear of other aerials and vertical masts, since these degrade the operation significantly.

Parts list

Qty. 25	M3×10 screw with countersunk head
Qty. 6	M3×16 screw with countersunk head
Qty. 3	M3×16 screw with cylinder head
Qty. 9	hexagonal nut M3
Qty. 16	flat spring washer M3
Qty. 8	solder tag for M3 screw
Qty. 12	self-locking bolt M3
Qty. 12	threaded bush M3
Qty. 4	cylinder-head screw for BNC socket
Qty. 6	aluminium rod (1) size: 1000×10×5 mm
Qty. 6	aluminium rod (2) size: 300×10×4 mm
Qty. 2	aluminium plate (3) and (4) size: 10×11×2 mm
Qty. 2	man-made fibre or perspex disc (5); (6) size: 40×40×8 mm
Qty. 1	length of PVC drainage tubing (7) dia: 40 mm; length: 300 mm
Qty. 1	coax cable 50 Ω, e.g., RG58C/U
Qty. 1	BNC socket with flange
Qty. 8	solder tag M3



IN-LINE RS-232 MONITOR

by A. Rigby

Serial links between computers and peripheral equipment based on the RS-232 standard are notoriously difficult to get going for the first time. Much of the frustration computer users suffer while connecting-up serial equipment is caused by their inability 'to see what is going on' on the data and handshaking lines. The small in-line signal monitor discussed here largely solves this awkward problem for almost any equipment sporting an RS-232 input or output.

Connections, computer ports and cables claimed to comply with the RS-232 standard are so common these days that the original application of this serial interface is often forgotten or not even known. In computer land, it is a generally accepted fact that virtually all 'non-standard' RS-232 links — even those of the so-called 'zero-modem' type — take a lot of valuable time to get operational. Not surprisingly, it is often desired to have a simple tool available for monitoring the activity of data and handshaking signals. Before describing the operation and construction of such a tool, it may be useful to give a brief recapitulation of the basic operation of the RS-232 interface itself.

Standard RS-232: OK as far as it goes

The signals available on a RS-232 connector, whether male or female, 9-pin or 25-pin, are in principle intended only to ensure correct transmission and reception of data from so-called DTE (data terminal equipment) to DCE (data communication equipment). A DTE is generally any data source, but it is usually a computer. A DCE is any device that converts data in a manner that allows this to be actually carried over some distance to a receiving system. The best known example of DCE is the telephone modem (*modulator/demodulator*).

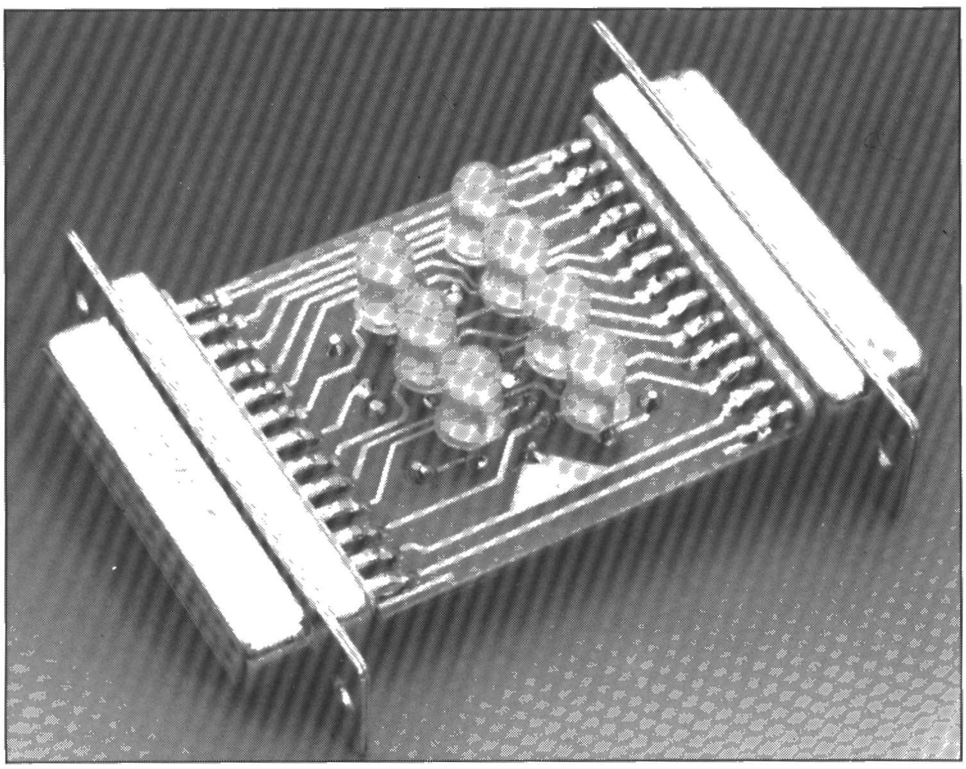
The RS-232 interface is specified such that DTE is linked to DCE by wires connected to pins with the same numbers on the connectors at both sides of the cable: DTE pin 1 goes to DCE pin 1, DTE pin 2 to DCE pin 2, etc. (see Fig. 1). Similarly, the signal functions are assigned such that data transmission is optimum on this multi-wire, but essentially simple-to-make, cable (see Table 1).

DTE-to-DTE = zero-modem

All was well with the RS-232 interface until, in the early seventies, someone decided to transfer files between two computers (DTE) by hooking up their RS-232 outlets. Such a connection between two DTE-type devices was not foreseen or, for that matter, specified or supported by the RS-232 standard, and obviates a good many handshaking signals. The so-called 'zero-modem' shown in Fig. 2 is known by now to virtually any PC user as a simple 6-wire cable (excluding ground which is not, strictly speaking carried over a wire) with one interconnection, 6—8, on each connector. In fact, the zero-modem is not a modem at all (whence its name): it merely acts as a single DCE 'seen' by both computers (DTE).

The other, even simpler, solution to DTE—DTE communication is the two-wire link, also shown in Fig. 2. Since this provides only handshaking to each individual computer, and not between the two of them, it may cause problems at relatively high data speeds. For most PCs and compatibles running the simple COPY COM1: instruction, the troubles typically start at 9,600 bits/s.

The attempts of some PC users to introduce handshaking for computer-to-com-



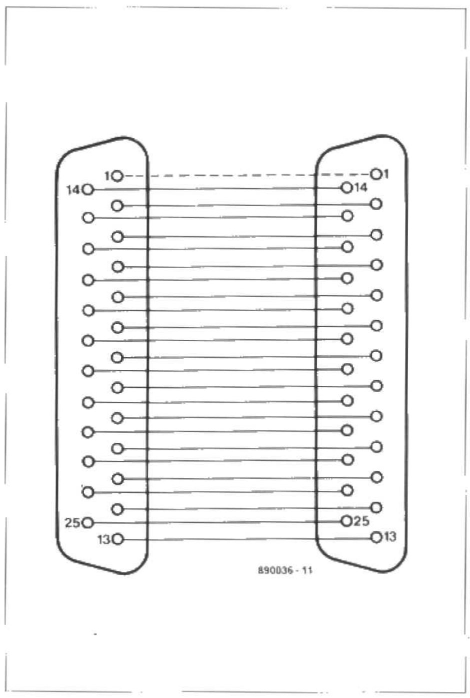


Fig. 1. Basic wiring dagram of a standard DTE—DCE 25-way RS-232 cable.

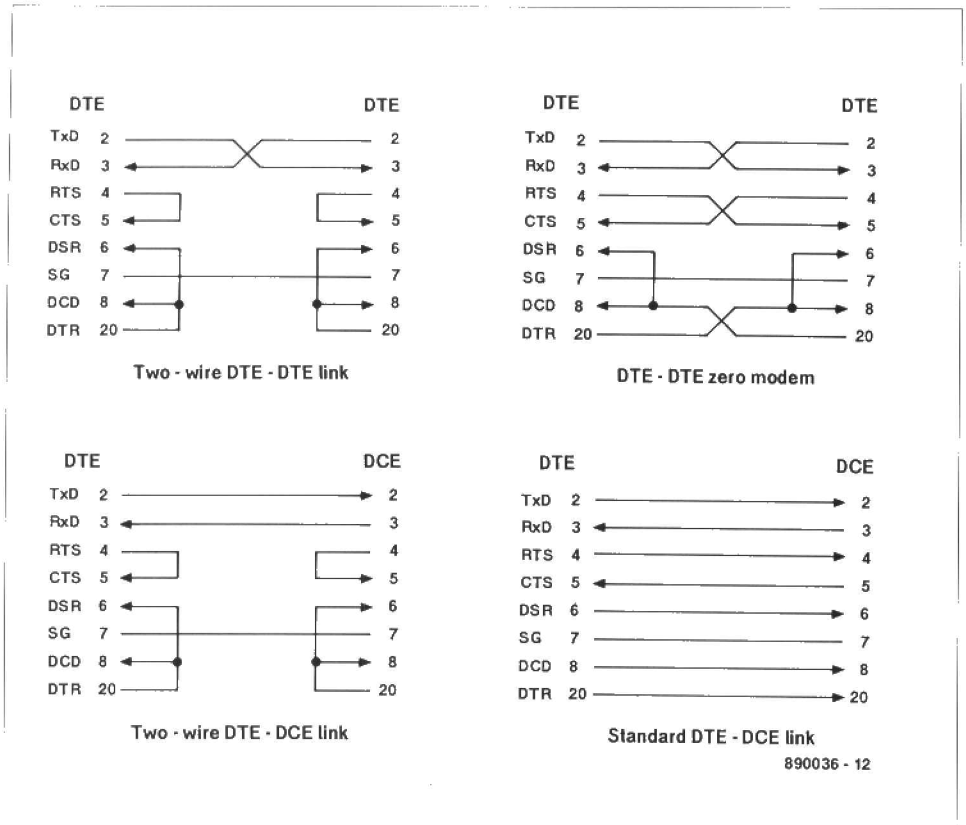


Fig. 2. Some commonly used RS-232 connections.

puter file transfer often rely on smart but essentially non-standard usage of the serial ports. Hence, these experiments are machine-specific and do not, in many cases, guarantee satisfactory results in other system configurations.

Bearing in mind that the RS-232 standard is still perfectly all right for everything it was originally designed for (bidirectional communication between DTE and DCE), it is fair to argue that a good deal of the compatibility problems experienced these days are caused by non-standard configurations and applications. By now, however, we seem to have accepted that the rapidly expanding use of computer-based communication has caused non-standard applications of the RS-232 interface to outnumber standard applications by far. So far, in fact, that the RS-232 interface is often unjustly critized for needless complexity while used in configurations it was never designed to handle.

Examples of RS-232-based, but definitely manufacturer-specific, serial interfaces include those on PC-ATs (the famous 9-pin connector), on Postscript laser printers that can 'talk back' to the computer, on equipment sending a non-symmetrical line voltage (down to simple digital drive with +5 V), and on a host of dot-matrix printers, intelligent modems, scanners and other digitizers, all commonly used in the PC environment. Time, therefore, for a simple tool that enables the 'communication expert' to quickly locate a problem if the serial link is no great shakes.

Circuit description

The circuit diagram of Fig. 3 shows that the signal indicator is built with a number of bi-colour LEDs, associated series resistors, two connectors, and a printed-circuit board to the design shown in Fig. 4. The tracks take all 25 pins of female 25-way D-connector K_1 at one side of the board direct to the male D-connector, K_2, at the other. Seven lines between K_1 and K_2 are 'tapped' to drive bi-colour LEDs that indicate the current logic level. The seven signals thus monitored are generally considered indispensable for correct data transfer via most RS-232 links.

As to the definition of the logic levels used on RS-232 datalines, remember that a logic one corresponds to a negative voltage, and a logic zero to a positive voltage (this does apply to the control and clock lines).

Construction

The printed-circuit board is small to ensure that the RS-232 monitor is a handy

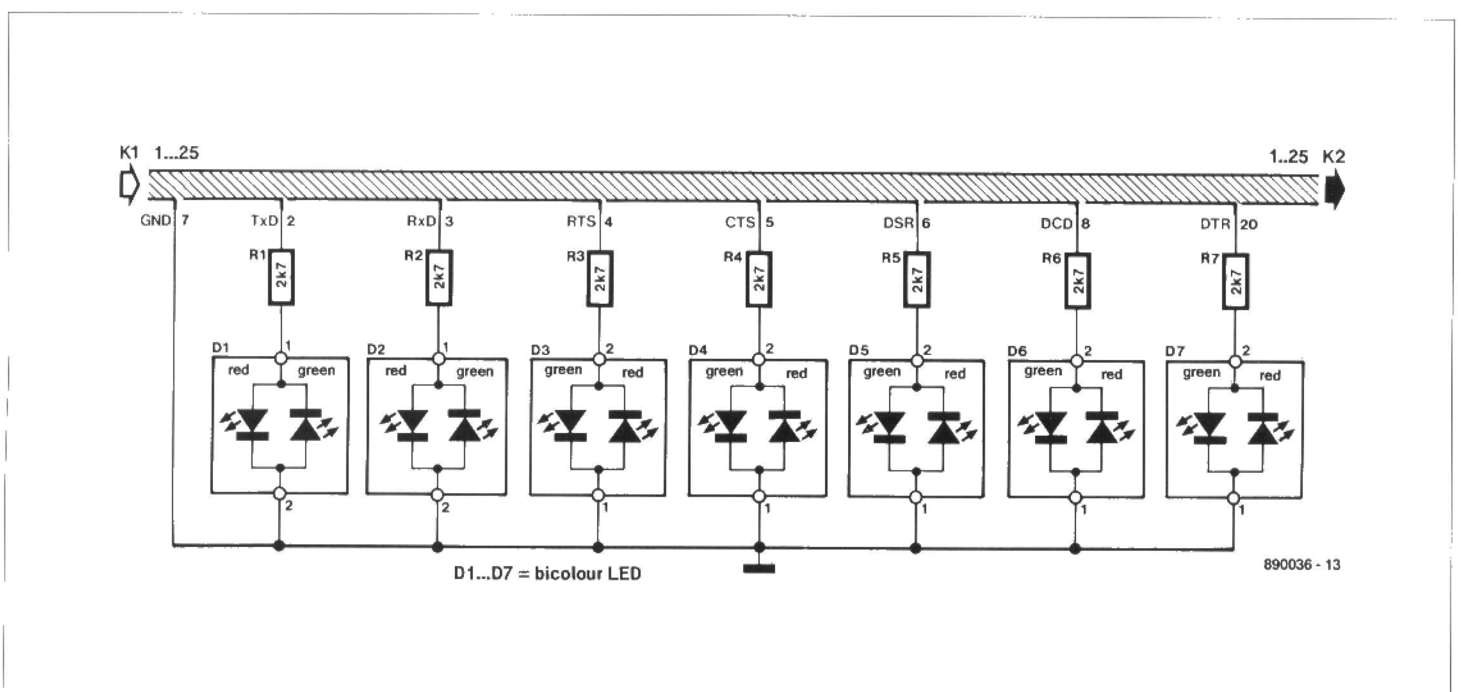


Fig. 3. Circuit diagram of the in-line RS-232 monitor. LEDs are used to indicate the status of the main signals carried via the serial link.

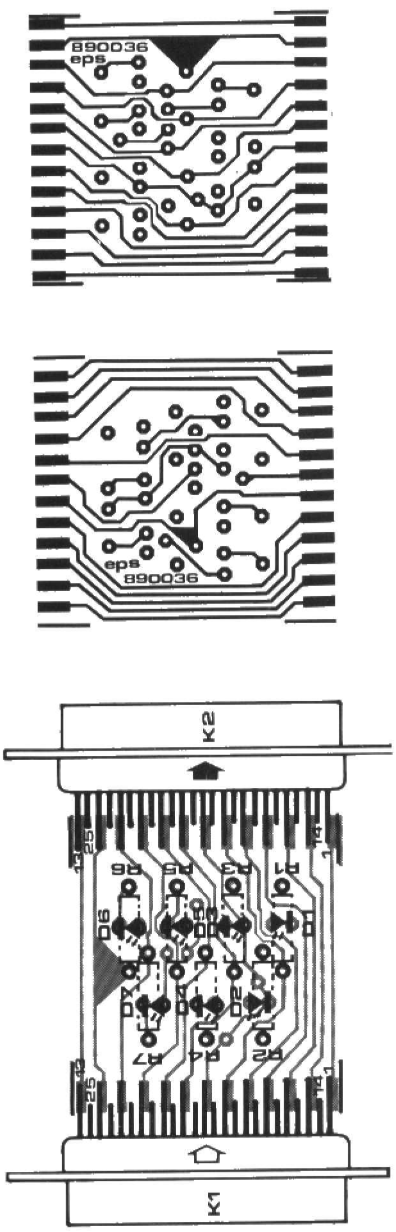


Fig. 4. Double-sided printed-circuit board for the RS-232 monitor.

Table 1.

Pin	Signal	Function	DTE	DCE
1	CG	chassis ground		
2	TxD	transmitted data	out	in
3	RxD	received data	in	out
4	RTS	request to send	out	in
5	CTS	clear to send	in	out
6	DSR	data set ready	in	out
7	SG	signal ground		
8	DCD	data carrier detect	in	out
9		positive test voltage		
10		negative test voltage		
11		not assigned		
12	SDCD	secondary DCD	in	out
13	SCTS	secondary CTS	in	out
14	STxD	secondary TxD	out	in
15	TxC	transmit clock (DCE)	in	out
16	SRxD	secondary RxD	in	out
17	RxC	receive clock	in	out
18		not assigned		
19	SRTS	secondary RTS	out	in
20	DTR	data terminal ready	out	in
21	SQ	signal quality detect	in	out
22	RI	ring indicator	in	out
23	SEL	speed selector DTE	in	out
24	TCK	speed selector DCE	out	in
25	BSY	data line busy	in	out

and rugged test device. The copper islands at the PCB edges are located in a manner that enables them to be soldered direct to the relevant pins of the 25-way female (K_1) and male (K_2) sub-D connectors (these are standard types with short, straight, pins, i.e., *not* special PCB-mount versions).

It is recommended to fit the two bi-colour LEDs for the RxD (received data) and TxD (transmitted data) reversed with respect to the other LEDs, so that a lit green LED always indicates a logic one.

The final appearance of the RS-232 monitor depends much on individual taste. The completed board may either be cast in an ABS moulding, covered by cut-to-size metal plates, or built into an enclosure made from the hoods supplied with the D-connectors. These hoods are modified and then glued together to form a compact casing.

Parts list

Resistors (±5%):
R_1...R_7 incl.=2K7

Semiconductors:
D_1...D_7 incl.= bi-colour LED

Miscellaneous:
K_1= female 25-way sub-D connector.
K_2= male 25-way sub-D connector.
PCB Type 890036 (see Readers Services page). Raster step one

Consortium to Investigate European Mobile Communications

A British-led four-nation consortium of telecommunications users and manufacturers has been contracted by the European Commission to investigate applications of broadband mobile communications.

The contract represents one of a number of programmes described as "application pilots", which form a critical part of the EC initiative on research into advanced communications for Europe—RACE. This involves users in specifying and trying out new communication services.

On behalf of its eight consortium partners from the United Kingdom, Belgium, Federal Germany and France,Cambridge Consultants Ltd, (CCL) is co-ordinating the RACE Mobile Application Pilots—RMAPS—project to assess the effectiveness and implications of providing advanced pan–European mobile services through four projects.

Success for First Aeronautical Digital Switch Trial

Ferranti Computer Systems has completed the world's first trial of a new digital transmission system.

The common ICAO Data Interchange Network—CIDIN—will replace the existing traditional telegraphic methods of switching aeronautical information, used by civil aviation authorities, with state-of-the-art high-speed, packet-switching computer communications techniques.

Current AFTN working is via relatively slow 50, 75, 100 and 300 bits/s transmission rates, which often requires manual intervention by operators, has no effective automatic error correction and recovery, no graphics transmission capability and is commonly conveyed on traditional landline or, in remote areas, via high frequency radio.

CIDIN, however, will provide higher transmission rates of 9600 bits/s on leased lines and up to 64 kbits/s on digital links, automatic error correction and recovery, a minimum of manual intervention and full graphics transmission capability for maps.

DIGITAL ECHO UNIT EG1000

design: ELV GmbH

The EG1000 echo/reverberation unit uses delta-modulation on a digital basis, and is capable of providing delay times of up to one second over a large bandwidth at an excellent signal-to-noise ratio.

A new, advanced, echo unit with superb technical specifications and a good price/performance ratio will no doubt be welcomed by a great many audio enthusiasts.

To begin with, however, it is perhaps useful to discuss briefly the difference between echo and reverberation, since these operations on audio signals are often confused.

General description

In an electronic echo system, intermediate signal storage works independently of the input sound, so that the delayed sound, which includes speech, is identical to the original. Reverberation systems, on the other hand, produce sounds with different delay times, so that speech is often reduced to noise or a continuous sound. The EG1000 echo unit enables reverb-like signals to be generated by selecting the REPEAT ECHO mode in conjunction with a delay setting between 80 and 100 ms. Each delay cycle then gives rise to a new, additional, delay, which produces an accumulation effect. An example:

80 + 80 → 160 + 80 → 240 ms + 80 ...

Tape-based echo units and reverberation springs are rapidly becoming obsolete. Here, too, digital technology has taken over. The EG1000 is such a digital unit for the following people and applications:

- professional and other musicians and electrophonics enthusiasts who want to give a special touch to vocals and music;
- film enthusiasts who want to do the same with their audio tracks;
- discotheques and clubs;
- connection to an existing home organ;
- sound-effect creation; post-dubbing, mixing and recording.

The EG1000 can introduce delays of 60 ms to 1 ms on audio signals, which are reproduced as a single or multiple echo (repeat option). The repeat mode allows even longer delays to be achieved.

The unit has two individually controlled inputs. Input 1 is accessible on the front panel for the connection of a dynamic microphone. Input 2 is accessible on the rear panel, and accepts signals from tape recorders, tuners, and musical instruments. This input has an additional switch, DELAY ON/OFF, which enables the signal to be fed to the output with or without echo. Input 1, on the other hand, is always subject to echo.

The two input levels are set with the aid of associated volume controls on the front panel. The total level setting is monitored by a LED functioning as a TRIP indicator, which should light rhythmically with the input signal, not continuously, to ensure optimum drive for the analogue-to-digital converter.

Like input 2, output socket 1, for connecting to a power amplifier, is accessible at the rear panel. A second output supplies the single-echo signal at full amplitude.

The echo delay time can be set to any value between 62.5 ms and 1 s by means of two controls: the 'coarse' selector, a three-way switch with settings 62.5, 250 and 125 ms, and the 'fine' control, a potentiometer that provides a multiplier between 1 and 4. A second potentiometer, marked DELAY LEVEL, sets the output amplitude of the echo signal relative to that of the original. Switch VIBRATO selects frequency modulation of the audio signal. The vibrato effect is not as effective with speech and composite music signals as it is with sounds of relatively long duration and fixed frequency as, for instance, produced by an electronic organ.

The echo mode is either single or multiple, as selected by the SINGLE/REPEAT switch on the front panel. In the repeat mode, the delayed signal is repeatedly fed back to the input, until it dies out. The feedback level depends on the setting of the DELAY LEVEL control. When this is set too high, the input may be overdriven.

The noise cancelling circuit can be actuated with the associated switch on the front panel. Noise cancelling should be used to ensure an acceptable signal-to-noise ratio for input sounds with a large dynamic range. The circuit is active only during the quiet periods of the input signal, so that the overall quality of the programme is not affected. In practice, old tape recordings may be considerably improved in this way.

Principle of operation

The audio signals to be processed are applied to a preamplifier that drives module IC_3, which converts the signal into digital, serial, data. To prevent overdriving, a LED connected to the module indicates the optimum drive level. Two dynamic random-access memories (RAMs) function as a shift register to achieve intermediate storage of the digitized input

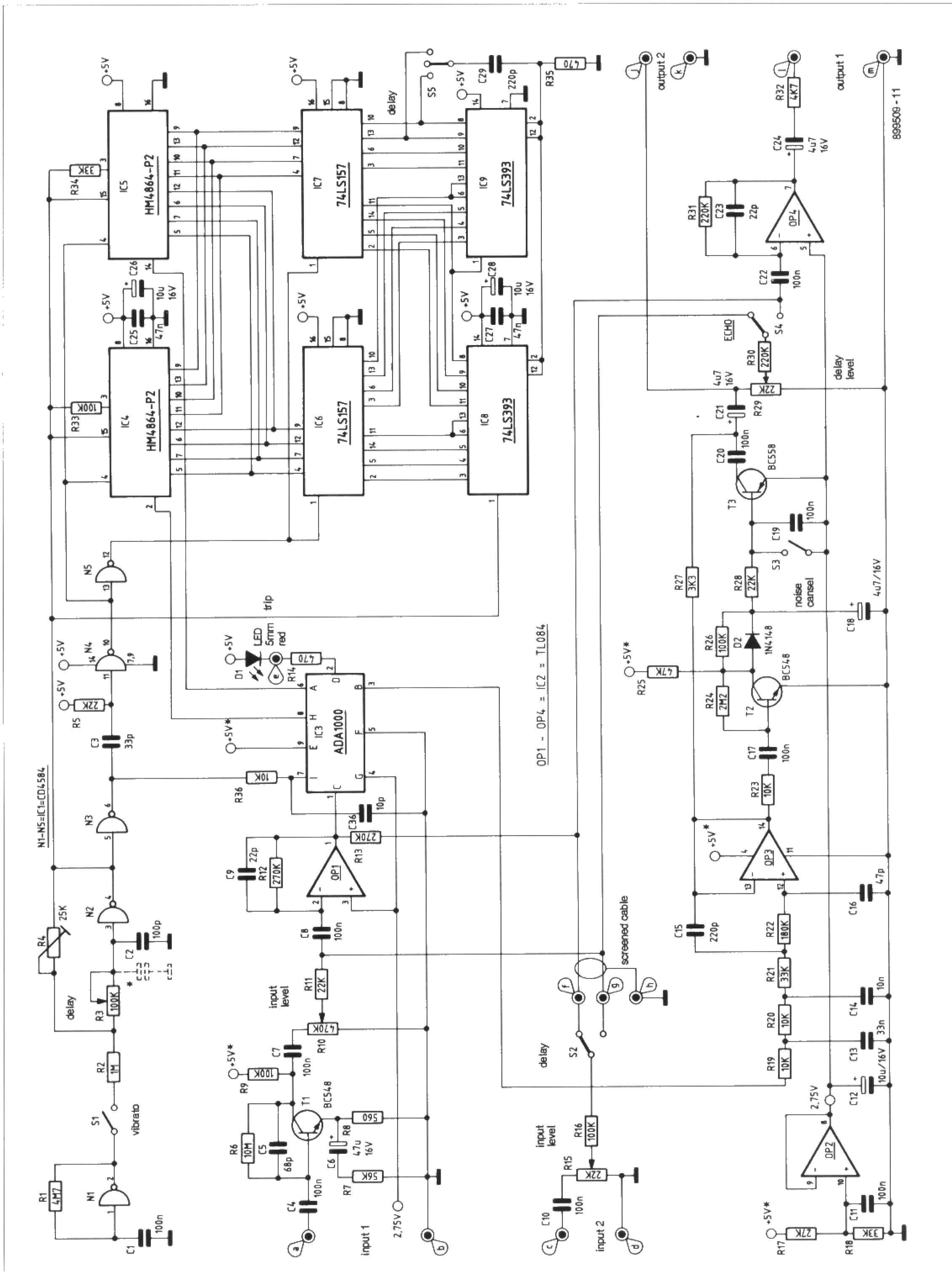
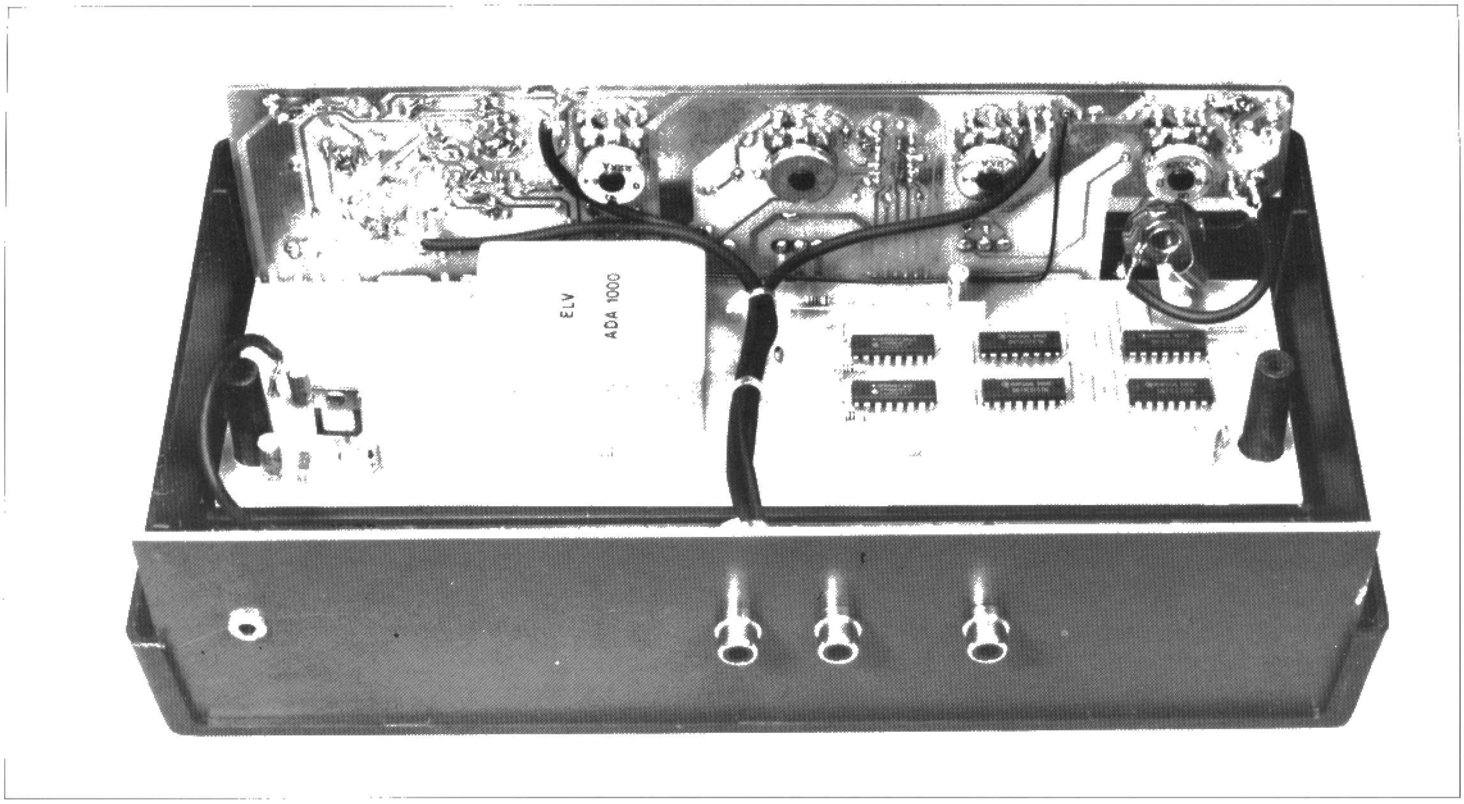


Fig. 1. Complete circuit diagram of the echo unit. Signal delays are achieved with a modular analogue-to-digital converter in conjunction with a 128 kBit RAM.



signal. Either RAM provides 65,536 memory locations, which are accessed by a pair of counters. Each address cycle starts with the removal of the previous data from the memory cell, so that new data can be written at the relevant location. The new data is not read out until all 131,072 addresses have been accessed. The delayed digital information is also fed back to the previously mentioned converter module, which converts it back to analogue. The converter thus has a double function: conversion from analogue to digital, and conversion from digital to analogue. The analogue output signal of the module is applied to an active low-pass filter that suppresses harmonics generated in the conversion processes.

The signal-to-noise ratio of the signal is improved by passing it through a switchable noise eliminator which is active during signal pauses only. The output signal of the noise filter is applied to an amplifier that supplies the output signal of the echo unit to an external power amplifier.

Circuit description

The circuit diagram of the EG1000 echo unit is given in Fig. 1. Signals applied to the microphone input (input 1) are amplified by T_1 to ensure an amplitude that allows mixing with the line signal (input 2). Mixing takes place at the input of opamp OP_1. Input 2 is connected either to OP_1 (delayed signal) or to OP_4 (not delayed), as selected with switch S_2. The non-delayed mixed signal of inputs 1 and 2 reaches OP_4 via R_{13}, and the delayed mixed signal via R_{29} (the DELAY LEVEL control), and R_{30}. The delayed signal at the output of OP_1 is also fed to A-D-A converter IC_3. Analogue input signals at chip terminal C are converted into digital data, which is available at terminal H. The input signal amplitude is monitored, and LED D_1 lights when a level of 75% of the maximum level is exceeded. The second function of the A-D-A module is the re-conversion to analogue of the serial digital data supplied by the RAM-based shift register.

The digital output data of the module is applied to cascaded RAMs IC_4 and IC_5, which are addressed by counters IC_8 and IC_9. The delayed digital data does not reach input terminal A of the module until all 131,072 memory locations have been addressed in succession.

The RAMs are addressed in two stages. Eight row and column addresses are subsequently applied. Multiplexers IC_6 and IC_7 convert the 16 bit counter output word into 2×8 format.

The clock oscillator of the echo unit is formed by N_2, whose output frequency lies between 130 and 520 kHz, as set with the DELAY potentiometer, R_4.

Gates N_3, N_4 and N_5 generate the control and timing signals required for the converter and the memory.

Gate N_1 is used as a low-frequency R-C oscillator that modulates N_2 when S_1 is closed. The result is frequency modulation required for the vibrato effect.

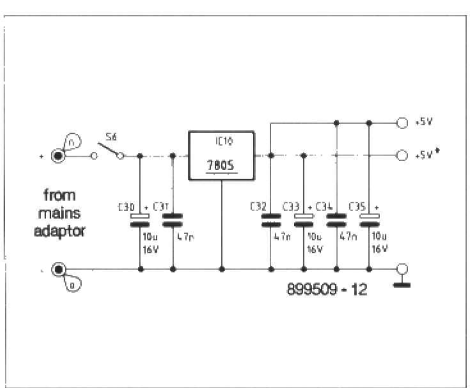


Fig. 2. Circuit diagram of the on-board voltage regulator.

The delayed audio signal at output terminal B of IC_3 is fed to active low-pass OP_3. This filter is dimensioned for a roll-off frequency of about 12 kHz to suppress harmonics in the output signal of the converter.

Transistor pair T_2-T_3 forms the noise eliminator. T_2 functions as a preamplifier, and T_3 as an electronic switch. During signal pauses, T_3 conducts and shunts off a large part of the noise signal via C_{20}.

The output signal of the EG1000 is supplied by OP_4, which guarantees enough drive for almost any power amplifier connected to output 1.

Capacitors C_5, C_9 and C_{23} provide some feedback to reject RF signals.

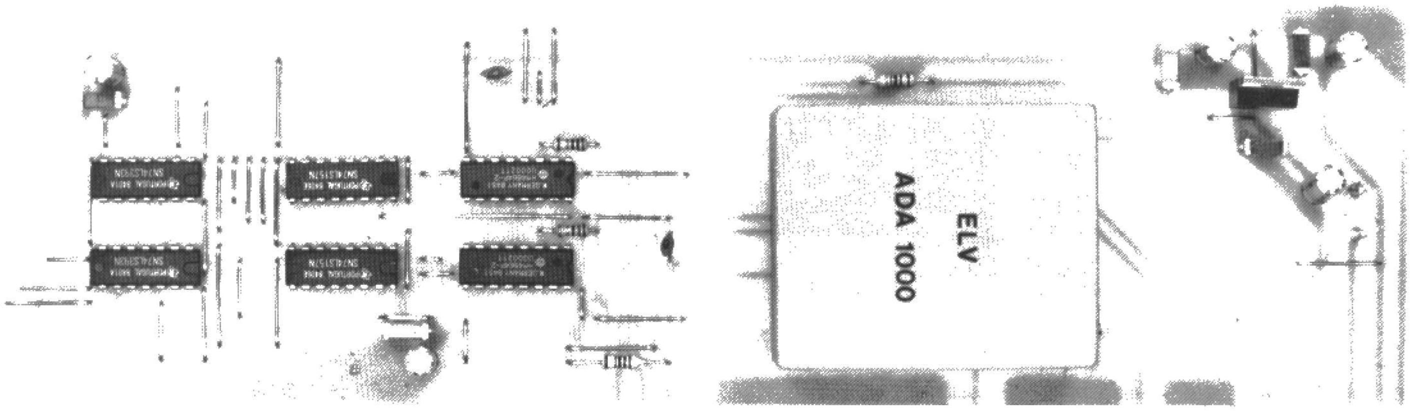
Opamp OP_2 provides a virtual ground rail at +2.75 V, which serves as the reference potential for OP_1, OP_3 and OP_4, eliminating a second supply voltage.

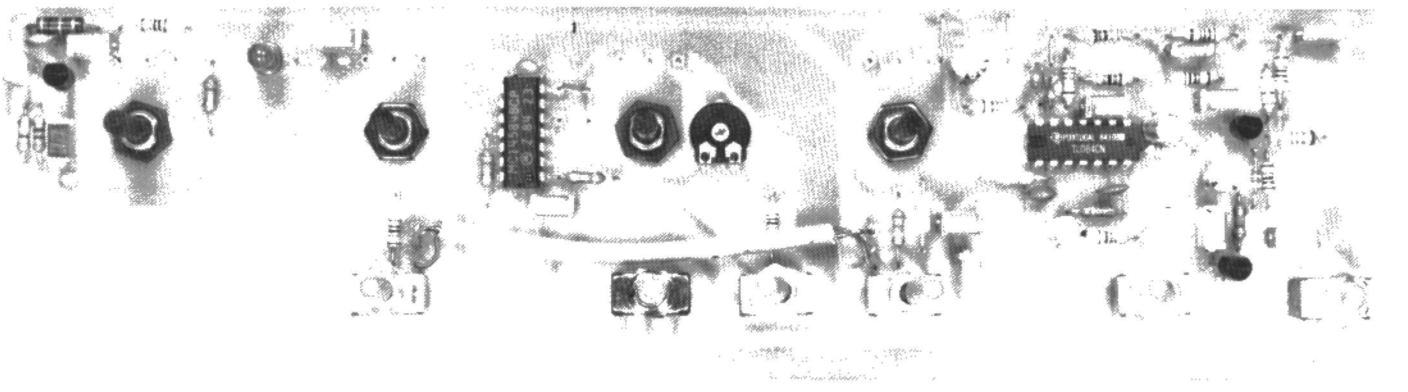
Switch S_5 selects the memory size and thus the delay time. Depending on the switch position, counters IC8 and IC_9 are reset at ¼ or ½ of the count range. When S_5 is open (centre contact), the full memory capacity is used, resulting in the maximum delay time.

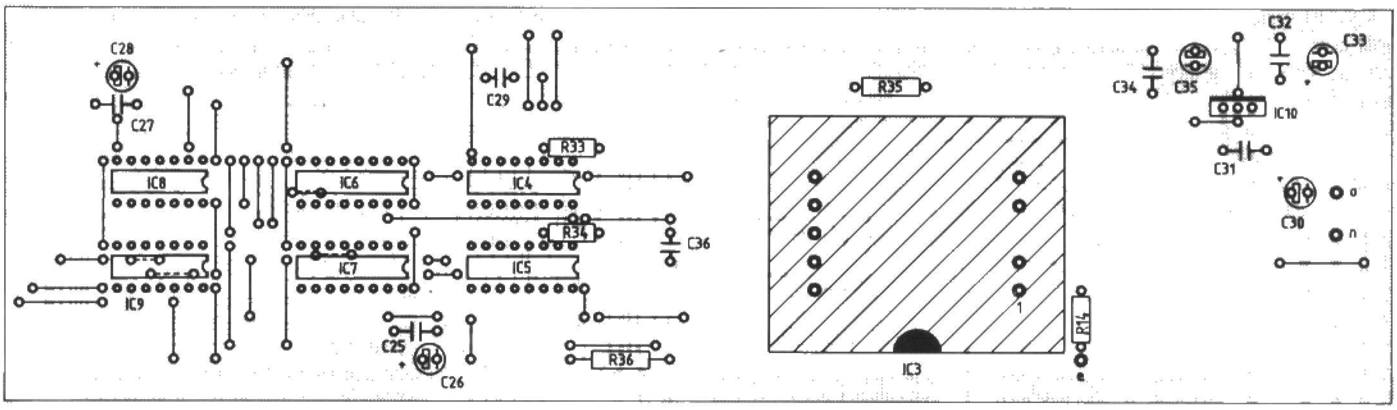
Construction

Although based on a fairly complex circuit of up-to-date design, the EG1000 is remarkably simple to build. This is mostly by virtue of the ready-made printed-circuit boards supplied with the kit, and the use of a pre-aligned A-D-A module.

It should be noted that the circuit is based on components that work with fast switching signals, and this requires great attention to be paid to the circuit board layout. This means that the echo unit







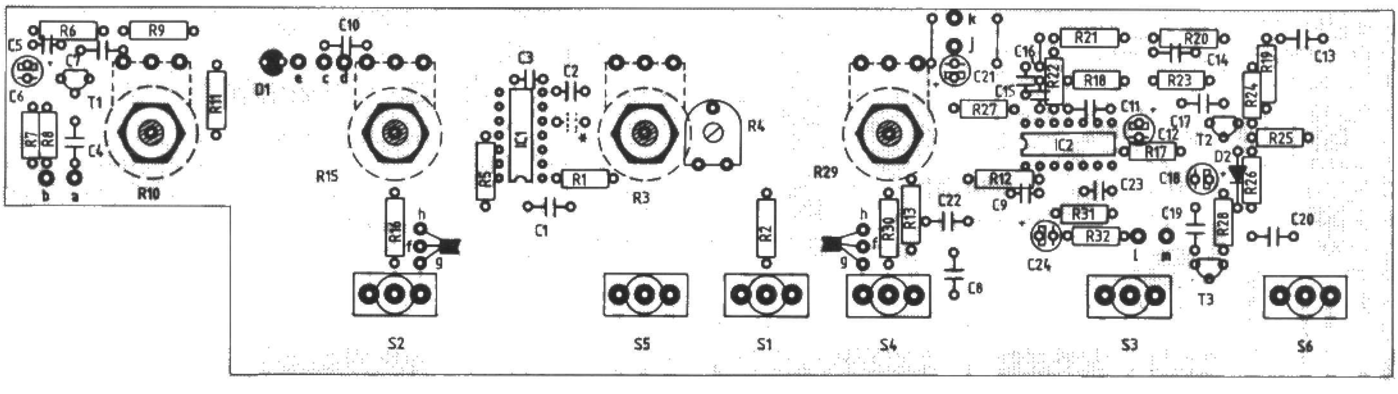


Fig. 3. Top views of the printed circuit boards, and their component mounting plans.

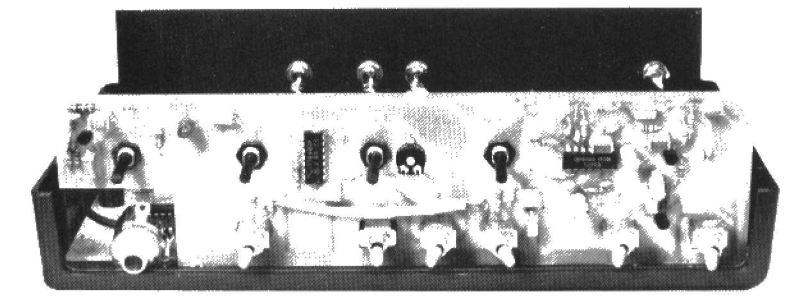
Parts list

Resistors:
R_1=4M7
R_2=1M0
R_3=100K linear potentiometer; spindle dia. 4 mm
R_4=25K preset H
R_5;R_{11}=22K
R_6=10M
R_7=56K
R_8=560R
R_9;R_{16}=100K
R_{10}=470K linear potentiometer; spindle dia.4 mm
R_{12};R_{13}=270K
R_{14}=470R
R_{15}=22K linear potentiometer; spindle dia. 4 mm
R_{17}=27K
R_{18};R_{21}=33K
R_{19};R_{20};R_{23}=10K
R_{22}=180K
R_{24}=2M2
R_{25}=47K
R_{26};R_{33}=100K
R_{27}=3K3
R_{28}=22K
R_{29}=22K linear potentiometer; spindle dia. 4 mm
R_{30};R_{31}=220K
R_{32}=4K7
R_{34}=33K
R_{35}=470R
R_{36}=10K

Capacitors:
C_1,C_4;C_7=100n
C_2=100p
C_3=33p
C_5=68p
C_6=47μ; 16 V
C_8;C_{10};C_{11}=100n
C_9;C_{23}=22p
C_{12}=10μ; 16 V
C_{13}=33n
C_{14}=10n
C_{15};C_{29}=220p
C_{16}=47p
C_{17}=100n
C_{18};C_{21};C_{24}=4μ7; 16 V
C_{19};C_{20};C_{22}=100n
C_{25};C_{27}=47n
C_{26};C_{28}=10μ; 16 V
C_{30};C_{33};C_{35}=10μ; 16 V
C_{31};C_{32};C_{34}=47n
C_{36}=10p

Semiconductors:
D_1= LED, 5 mm dia.
D_2=1N4148
IC_1=CD4584
IC_2=TL084
IC_3=ADA1000
IC_4;IC_5=HM4864-P2
IC_6;IC_7=74LS157
IC_8;IC_9=74LS293
IC_{10}=7805
T_1;T_2=BC558

Miscellaneous:
5 off miniature SPDT switch.
1 off miniature SPDT switch with centre-off position.
1 off 6.3 mm jack socket.
1 off 3.5 mm jack socket.
3 off phono socket.
10 off solder terminals.
60 cm screened wire.
20 cm light-duty insulated wire 2×0.4 mm^2.
80 cm silvered wire.
10 cm stereo screened wire.



View of the vertically mounted controls board with the front-panel removed.

meets the required specifications only when constructed on the boards shown in Fig. 3. Construction itself is fairly simple for those who have some experience in building medium-sized electronics projects. Careful soldering is a must, however, and some precautions are necessary in this respect as will be discussed below.

The EG1000 has an internal regulator that takes its input current from a d.c. mains adapter with 9–15 V output. Current consumption is modest at about 50 mA. Since the EG1000 has no mains-carrying wires or connections inside, it is safely used with its enclosure opened.

The construction is carried out in the following order.

Start with populating the boards by fitting and soldering the components. It is recommended to fit the low-profile parts, such as the resistors and wire links, first, followed by the higher parts. The integrated circuits aré fitted last.

The converter module must be mounted with utmost care, and its pins are to be soldered as quickly as possible. Pause for 5 seconds or so between successive solder actions to prevent the module overheating inside. The points marked 'f', 'g' and 'h' in the circuit diagram are found back on the front-panel board. These terminals are connected to the main board by approximately 10 cm long screened wires. Terminal 'h' is connected to the screening, and terminals 'g' and 'f' to the centre cores.

Following the completion of the printed-circuit boards, all solder junctions must be inspected carefully. This is particularly necessary in the area around IC_4 to IC_9, where the track density is relatively high. In case of doubt, use a magnifying glass to look for short-circuits created by excess solder tin.

The completed boards are then joined at right angles, with the front-panel board protruding about 2 mm under the track side of the main board. Next, install the screened cables to and from the input and output sockets. Be sure to keep these cables as short as possible.

The power supply wires are connected to the 3.5 mm jack socket on the rear panel by light-duty wire.

Simple-to-adjust!

The A-D-A converter in the EG1000 has carefully selected components inside, and has been tested and aligned by the manufacturer. This reduces the setting-up procedure of the echo unit to the adjustment of a single preset, R_4.

Before power is applied, turn R_4 fully clock-wise, after which the following controls are set:

- R_3 to position '1'
- R_{29} to position '100%'
- R_{10} and R_{15} to position '0%'
- switch S_4 to 'repeat'
- switch S_3 to 'off'

Then, connect the previously inspected echo unit to an amplifier, and adjust preset R_4 until a high-pitch tone is heard. Finally, R_4 is re-adjusted until the tone is inaudible when R_{29} is quickly set to 0% and back to 100%. This concludes the setting-up of the EG1000.

Note: we regret that photocopies and/or films of PCB Types ELV38255 and ELV38256 can not be supplied through the Readers Services.

A complete kit of parts for the EG1000 echo unit, which is designed in West-Germany, is available from the designers' exclusive worldwide distributors (regrettably not in the USA and Canada):

ELV France
B.P. 40
F-57480 Sierck-les-Bains
FRANCE
Telephone: +33 82827213
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SEMICONDUCTOR DIODES

by T. Wigmore

Although many readers know perfectly well what a diode is, it does no harm to repeat its definition here: it is any electronic device that has only two electrodes. There are two types of diode: thermionic and semiconductor. The present article will discuss semiconductor types only.

A semiconductor diode is basically a p–n junction, that is, a junction of n-type and p-type semiconductor material, currently usually silicon. An ideal junction of this nature, forgetting for the moment special types, such as zener diodes and varactors, behaves either as a short-circuit or as an infinite resistance, depending on the polarity of the applied voltage. Such a diode would possess differential resistance, r_d, and d.c. resistance, R_d, only. Unfortunately, ideal components do not exist and in a practical diode other parameters, such as bulk resistance, R_b; junction capacitance, C_j; diffusion capacitance, C_d; case capacitance, C_c; and terminal inductance, L, also affect its behaviour. These parameters are shown diagrammatically in Fig. 1.

The deviation of a practical from an

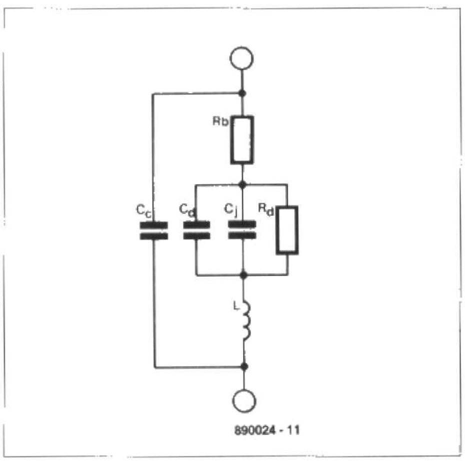


Fig. 1. Equivalent circuit diagram of a typical small-signal or switching diode.

ideal diode may be seen from the typical diode characteristic in Fig. 2. In the forward bias region, R_d is fairly large until the threshold voltage is reached, after which it is small. In the cut-off region (note the different voltage scale), only a small (leakage) current flows in the diode until breakdown occurs, after which, except in the case of zener diodes, the diode is destroyed.

Direct voltage

When the voltage applied across the diode is direct or alternates very slowly, only R_b and R_d affect the behaviour of the diode: the other parameters in Fig. 1 may be ignored.

The diode characteristic is then a function of the two resistances only. Since we can

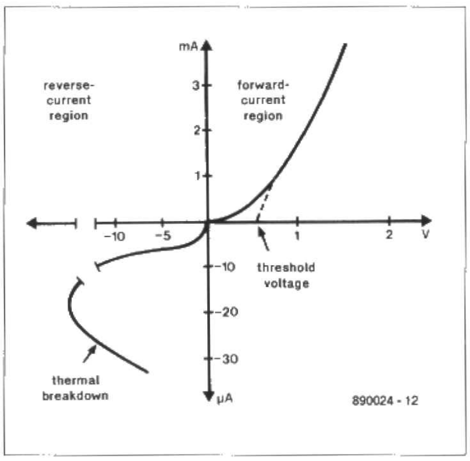


Fig. 2. Typical diode characteristic. Note the different scale of the $-x$ and the $+x$ axes.

not deal with the derivation of the formulas for these resistances in this article, we can only say that the threshold voltage in silicon diodes is 0.5–0.8 V and that in germanium diodes, 0.2–0.4 V. Once the threshold voltage is reached, the current would rise fast and linearly, were it not for the bulk resistance, which tends to impede the current, as can be seen in Fig. 3.

In the reverse bias region, R_b is of little significance, since it is negligibly small compared with the conductance, G_d.

The characteristic of a germanium diode is flatter than that of a silicon diode, both in the forward and in the reverse bias

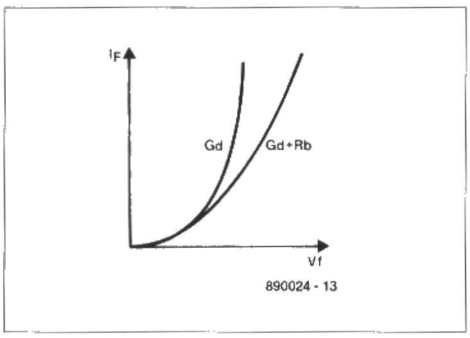


Fig. 3. Current vs applied voltage characteristic in the forward bias region with and without the effect of bulk resistance.

region.

Alternating voltage

When an alternating voltage is applied across the diode, the various capacitances inherent in the diode (see Fig. 1) become the dominant parameters. Even at low-frequency voltages, these capacitances may make the diode unsuitable for certain applications.

The relation between applied voltage, time and the consequent current through the diode is shown in Fig. 4.

The junction capacitance is important for the behaviour of the diode in the reverse bias direction, when a dense space charge exists at the p–n junction. At the instant the diode switches to reverse bias operation, the current through the junction capa-

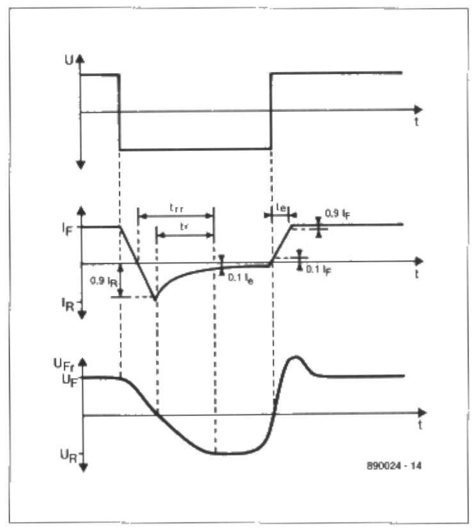


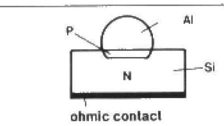
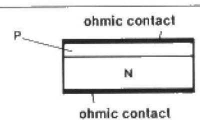
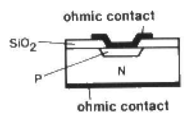
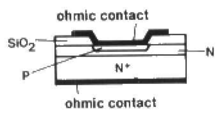
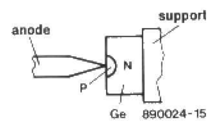
Fig. 4. From the top: the applied voltage, the current vs time curve and the voltage vs time curve.

citance changes polarity (I_F to I_R), and rapidly declines to a very low value (the leakage current, which is of the order of a few nanoamperes). The time it takes I_R to fall from 90% to 10% of the value of I_F is called the recovery time, t_r.

When the voltage rises, C_j decreases exponentially, since the width of the space charge region increases.

At zero crossings of the applied voltage, the diffusion capacitance, C_d, also affects the switching times, since the char

Table 1

Type of diode	Construction	Properties	Applications
Alloyed junction		Large cross-sectional area of barrier layer; large capacitances; high currents; large tolerances	Power diodes; zener diodes up to 10 V
Diffused junction		Large cross-sectional area of barrier layer possible; wide range of capacitances	Power diodes; zener diodes above 10 V
Planar		As diffused junction types but with much tighter tolerances; small dimensions and capacitances possible; good HF characteristics	General purpose; zener diodes; varactors; p-i-n diodes; Schottky diodes; HF diodes; switching diodes
Planar epitaxial		As planar types but with very low forward resistance and very short recovery times	
Point-contact		Very small capacitances; only small currents permissible; good HF characteristics	General purpose (low reverse bias and low forward currents); HF (up to VHF region); switching diodes

ge carriers in the semiconductor material have a certain inertia and act as short-term memories, particularly when a conducting diode is switched rapidly to reverse bias operation. It is, therefore, a requirement of rectifier and switching diodes that their diffusion and junction capacitances are minimal.

At frequencies below about 100 MHz, the case capacitance and terminal inductance have but little effect, but as the frequency rises they become more and more influential and must, therefore, be included in any computations.

Types of construction

The construction, properties and applications of five types of diode are shown in Table 1.

Included in the table is the germanium point-contact diode in which, because of the very small contact area between anode (point) and the n-type germanium, the junction capacitance is very small (<1 pF) so that it is eminently suitable for high-frequency and fast-switching applications. The use of gold-gallium anodes allows switching times shorter than 1 nanosecond to be achieved. Also, its forward bias is smaller than that of silicon diodes. Against these advantages, it can not cope with currents in excess of about 10 mA.

Silicon point-contact diodes with similar advantageous properties also exist, but because of their high vulnerability to overloads they are not of great importance and are used only in very special applications.

Germanium junction diodes have been superseded almost completely by silicon junction diodes and are nowadays used only where low forward bias is vital.

Silicon junction diodes are produced principally by one of three methods. In the **alloy process**, the basic material is an n-type wafer of silicon doped with antimony into which an aluminium ball is inserted at high temperature. During the solidification process a sharply defined n-p region is formed owing to the different fusion points of the materials and the diffusion of Si atoms in the aluminium. Because of the large area of the junction, this technique ensures that large forward currents are possible, although the device parameters are subject to wide tolerances.

These tolerances are much smaller in the **diffused junction process**. In this, a wafer of n-type silicon with a very smooth surface is heated to 1300 °C in a diffusion oven after which its surface is changed to n^+ by a P_2O_5 dopant. Subsequently, the doping layer is removed from one side of the wafer after which this is doped with boron to make it p-type.

The wafer is then provided at both sides with a terminal alloy after which it is sliced into small discs.

The cross-sectional area, and thus the ensuing capacitance, may be given a fairly wide range of values. The diffused junction process is particularly suitable for manufacturing power diodes and varactors.

Planar diodes are produced by a quite different technique. In this, a layer of silicon dioxide, SiO_2, is thermally grown on the surface of a silicon substrate. Photolithography is used to etch holes in the oxide layer, which then acts as a mask for the diffusion of boron impurities to produce a p-type region. The crystal is then cut into small slices. This technique guarantees small dimensions, small capacitances and precise reproducibility.

Planar epitaxial diodes have an additional n^+ doped layer at the back which makes them extremely low-ohmic in forward bias operation.

Schottky diodes are planar epitaxial types without boron doping. Instead, they have a metal contact sintered direct on to the n-type substrate, which (because of the Schottky effect) acts as a p-type semiconductor. This has the advantage of greater hole mobility and, consequently, a smaller diffusion capacitance and shorter storage and switching times (about 100 picaseconds). Figure 5 compares the rectification of a 30 MHz signal in a Schottky diode and in a general-purpose diode.

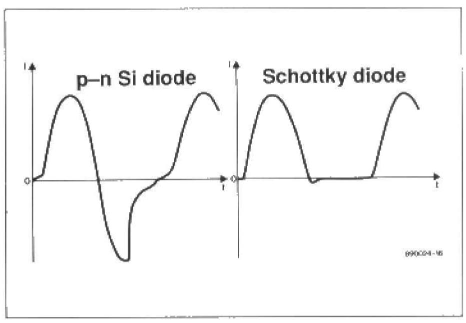


Fig. 5. The Schottky diode has definite advantages over a general-purpose diode for the rectification of a 30 MHz signal.

Practical diodes

After our short incursion into semiconductor theory, we shall now look at some practical diodes.

Small-signal diodes

The most popular small-signal diode is the 1N4148. Although this has been around for about 15 years and costs next to nothing, it has some very useful properties. With a parallel capacitance of not greater than 4 pF and a recovery time of 4–8 nanoseconds, it is eminently suitable for

use in h.f. circuits. Its family includes the 1N4149; 1N4446–1N4449; 1N914A; 1N914B; 1N916A and 1N916B, all with similar characteristics. A serious drawback of these diodes is their low forward current (max. 150 mA). Their reverse bias is of the order of 75 V and their dissipation around 440 mW. They are produced by the planar epitaxial technique.

In applications where a low voltage drop across the diode is required, the Schottky types BAT81–83 (switching time <1 ns) or BAT85–86 (switching time <4 ns) are used nowadays, where in the past germanium diode Type AA119 would have been used. The Schottky types have a lower voltage drop (<400 mV), but their reverse bias of 40–60 V is lower than that of the AA119.

Freewheeling and rectifier diodes

For mains voltage rectification at currents below 1A, the most suitable diodes are found in the 1N4001–4007 series. Their reverse voltage, depending on type, ranges from 50 to 1000 V. Apart from the fact that all diodes in the series are easily available, and at low prices, they can withstand short peak currents of up to 30 A.

For forward currents of up to 3 A, it is best to use one of the types in the 1N5400–5406 series, which withstand short peak currents of up to 200 A.

Both series are manufactured by the planar technique.

As an aside, a full-wave rectifier configuration using four discrete diodes is still cheaper than a proprietary bridge type.

Fast freewheeling and rectifier diodes

For operation at frequencies above 50 Hz, the diodes discussed above are too slow, and fast-recovery Types 1N4933–4937 should be used. These are similar to members of the 1N4001–4005 series, but have recovery times of 100–150 ns. These times guarantee satisfactory operation up to about 250 kHz. They are typically used in switch-mode power supplies.

Still faster are the BYV36A–36E series (reverse bias 200–1000 V; t_r <100 ns); the BYV26/50–26/200 (1 A types) and the BYV27/50–27/200 (2 A types). The latter two series, all planar epitaxial types, offer recovery times of not greater than 25 ns.

High-voltage diodes

High-voltage diodes are often encountered as rectifiers in cascode circuits. Their reverse bias is high—in the BY505: 2 kV and up to 24 kV in the BY741.

Diodes with low leakage current

Diodes with very low leakage current are very hard to come by. Fortunately, they may often be replaced by good Schottky types or, if really necessary, by a Type BF256B field-effect transistor of which the drain and source terminals have been interconnected.

Table 2

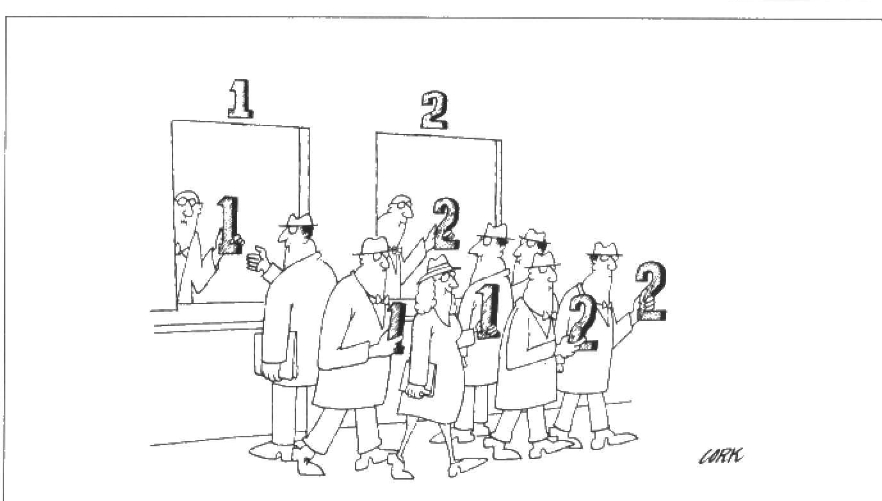
Type	Typical parameters	Applications
1N4148	Low forward current (200 mA; 400 mA max); fast (4 ns); inexpensive	Standard diode for small-signal and switching operation at low currents; free-wheeling diode for small relays
BAT85	Low forward current; fast; inexpensive	Schottky equivalent of 1N4148; used in inductance and millivolt meters
1N400X	Medium forward current (1 A); relatively slow; high peak currents up to 30 A	Low-frequency rectifier; freewheeling diode; suitable for mains operation
1N493X	Similar to 1N400X but faster (150 ns); 1N4937 suitable for mains operation	Fast rectifier; used in Elektor Electronics digital train decoder circuit
1N540X	Medium forward current (3 A); otherwise as 1N4001	Medium power rectifier
BYV27	Very fast switching diode (25 ns); medium forward current (2 A); low reverse bias	Freewheeling diode in stepper motor circuits; used in h.f. neon tube dimmers
BYV26	Similar to BYV27 but at higher voltage and lower current (1 A)	Used in h.f. neon tube dimmers
BYV36	Similar to BYV26 but slower	
BYV79	Fast switching diode at high currents (14 A)	Control circuits for radio control; used in 28 V converters
BYV19	Schottky rectifier at high currents (10 A)	Used in battery chargers

Fast power diodes

Fast power diodes are normally found in power supplies whose primary circuits are clocked and in motor control circuits. Suppressor diodes for operation at very high currents, such as the BZW86X (12–85 V at 250–1000 A; dissipation 25 kW) are not readily available and naturally tend to be very expensive.

At lower powers, the BYV79 or the Schottky BYV19 may be used. The BYV79 is particularly suitablefor use as a freewheeling diode. It can handle currents of up to 14 A, has a reverse bias, dependent on version, of up to 20 V. Unfortunately, it is not very fast (recovery time <50 ns) and has a voltage drop of 0.85 V at 10 A.

Where these aspects are important, it is better to use the Schottky version. This is not able to handle such large currents (up to 10 A), but its voltage drop of 0.6 V is significantly lower. Furthermore, its recovery time is only a fourth of that of the BYV79.



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Simplified time-signal receiver	87513	7.40	1.11
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LFA-150 — a fast power amplifier	880092-1	8.45	1.27
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Autonomous I/O controller	880162	4.50	0.68
	880163	4.85	0.73
	880184	15.30	2.30
Pitch control for CD players	880165	11.50	1.73

	No.	Price (£)	VAT (£)
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MIDI control unit	880178-1	9.05	1.36
	880178-2	6.65	1.00
Low-budget capacitance meter	UPB-S1	1.95	0.29
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The digital train	87291-1	4.20	0.63
Touch key organ	886077	10.05	1.16
Car service module	86765	3.60	0.54
	886126	4.10	0.62
VHF receiver	886127	7.45	1.12
Dark-room timer	886100	Not available	
MARCH 1989			
Diesel sound generator	880006	Not available	
VHF/UHF wide-band amplifiers	880186	Not available	
Power line modem	880189	6.10	0.92
ATN Filmnet decoder	890002	Not available	
Centronics buffer	890007-1	19.60	2.94
	890007-2	2.15	0.33
	890007-3	8.35	1.26
APRIL 1989			
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Multi-point IR control	890019-1	3.45	0.52
	890019-2	4.05	0.61
Video recording amplifier	889502	See advert on pp.8-9	
MAY 1989			
RDS decoder	880209	Not available	
Digital model train (4)	87291-4	5.25	0.79
Analogue Multimeter	890035	12.50	1.88
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DTMF system decoder	890060	6.50	0.98
Transistor tester	896029	Not available	
Sine wave converter	UPB-S1	1.95	0.29
Super-VHS to RGB converter		See advert on pp. 8-9	
JUNE 1989			
8-digit frequency meter	880128	11.50	1.73
In-line RS232 monitor	890036	Not available	
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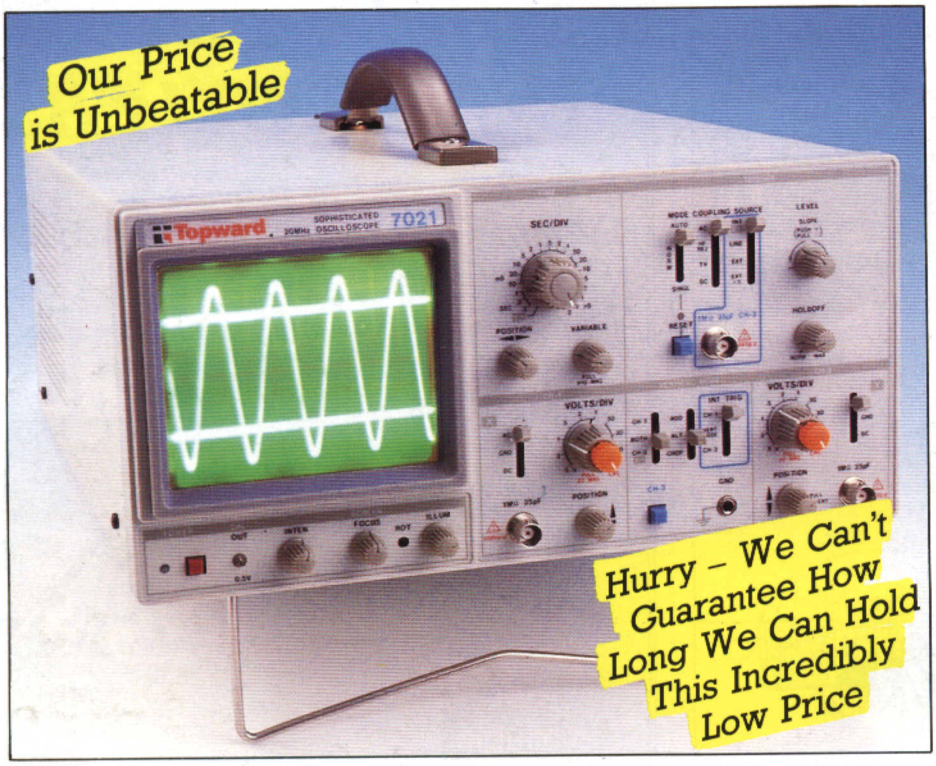

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